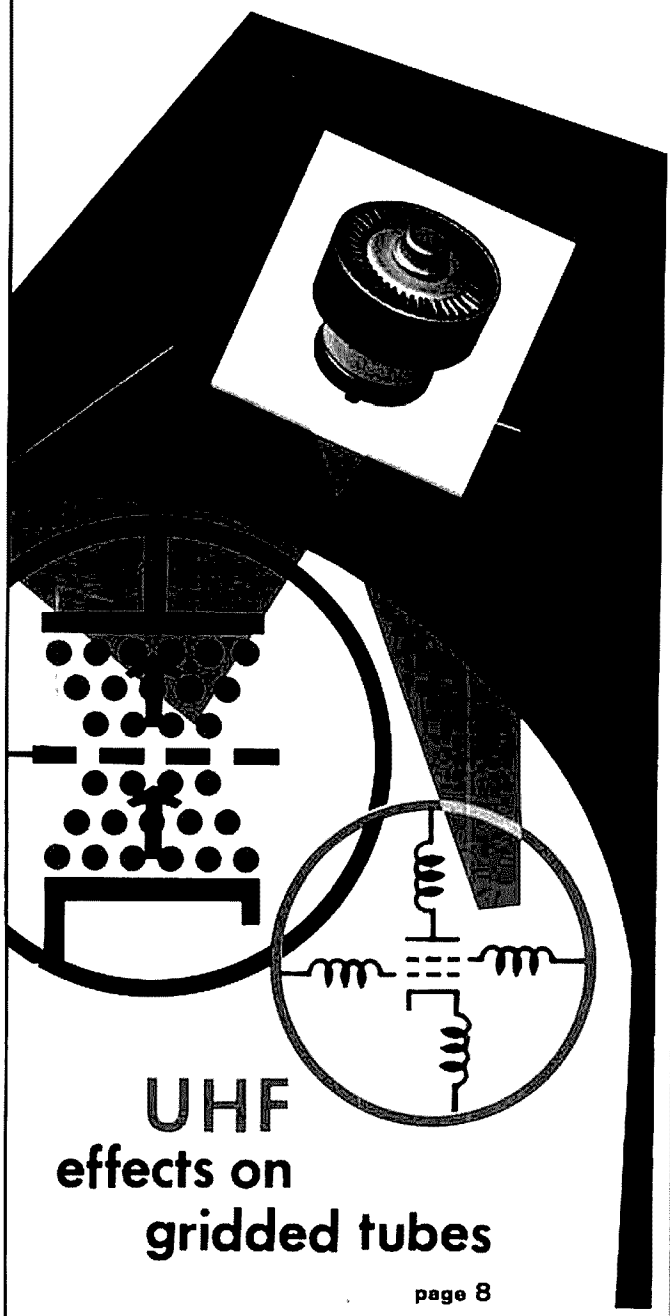
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JANUARY 1969



UHF effects on gridded tubes

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staff

editor
James R. Fisk, W1DTY

roving editor
Forest H. Belt

vhf editor
Robert M. Brown, W9HBF

associate editors
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Nicholas D. Skeer, K1PSR
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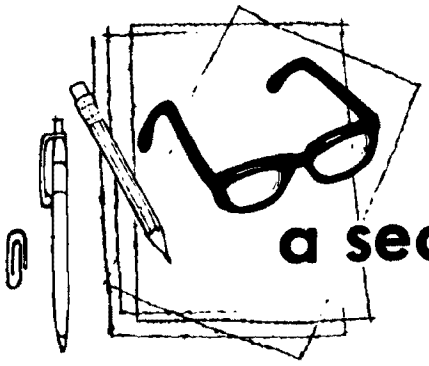
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a second look

by jim fisk

At one time or another I suppose most of us have wondered why our semiconductors are numbered as they are with 1N, 2N and 3N numbers. To the newcomer, it must look like a meaningless bunch of random numbers—and it isn't a great deal more to the oldtimer! Receiving-tube numbers follow certain guidelines, as do many transmitting tubes, so what happened to solid state? To say that it is a numbering system that grew on us isn't too far from wrong—but that's getting ahead of the story.

During the early days of radio, each vacuum tube manufacturer numbered his tubes as he saw fit. When there were only a few tube types available—with very little difference in their characteristics—this was no great problem. In fact, if you look in any of the old radio books, you'll see that none of the circuits specify what tube to use; it didn't make that much difference.

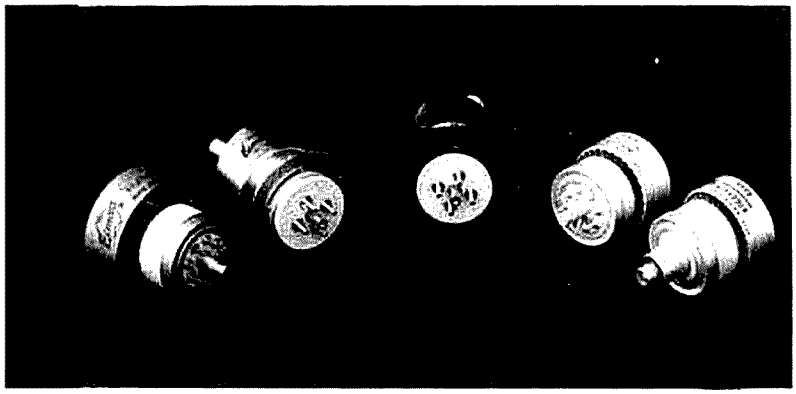
However, with the invention of the tetrode and pentode, and the introduction of tubes with vastly different characteristics, the problem became pretty sticky. The replacement problem was particularly bad, so in 1933 the industry voluntarily adopted a numbering system that called for a number, a letter and another number. The first number denoted the filament voltage range, the middle letter was a serial designation and the last number indicated the number of useful elements for which terminals were provided (including internal shield and shell connections). This is pretty much the system still in use today.

Transmitting tubes didn't run into the standardization bugbear until 1942. With the war, and the great number of tube types being manufactured, a standard number-letter-number system was adopted for transmitting and special-purpose tubes. Examples of devices assigned number from this system are the 1N21, 2C39, 2E26 and 3E29. The first number indicated the power rating of the heater: 1 for zero power, 2 for up to 10 watts, 3 for up to 20 watts and so on. The letter indicated the structure or function: B for diodes, C for triodes, D for tetrodes, E for pentodes, N for crystal diodes and rectifiers, etcetera. The final number was a serial designation, started at 21 to avoid conflict with the receiving-tube system.

Except for the use of 1N for semiconductor diodes, this system was scrapped in 1946 for a purely numerical system starting at 5500. Although several manufacturers wanted a numbering system for diodes (and later transistors) that told the user more about the device than a simple serial system, none was ever agreed upon. In the meantime, the system we use today kept growing. The first digit came to indicate the number of elements minus one—thus a 1N34A is a diode, a 2N706 is a triode and a 3N159 is a tetrode; the N indicates a solid-state device, and the last number is the order of registration.

This brings up another question—who decides precisely what number will be assigned to a particular transistor? The answer is an industry sponsored committee that registers all new transistors (and vacuum tubes), assigning the next open number in the system. If the system doesn't make much sense to you, don't feel too badly; a number of professional groups have tried to put some meaning into it, but without a great deal of success. The big argument against any new system at this point in time concerns the great number of solid-state devices already on the books—making any new numbering system impractical.

Jim Fisk, W1DTY
Editor



vhf/uhf effects in gridded tubes

Here is a rundown on the characteristics you should understand to get the most out of gridded tubes at uhf

Bob Sutherland, W6UOV, Power Grid Development Lab, Eimac Division of Varian, San Carlos, California 94070

The technically-minded amateur has known for some time that a vacuum tube becomes progressively less effective as the frequency of operation is increased. Amplifiers require greater driving power, output power drops off, and the manufacturer may be obliged to derate the tube at the high-frequency end of the operable range. If the frequency is raised high enough, the gain of the amplifier will drop to an unusable level. Upon further increase in frequency, the gain will drop to unity or less. At the same time this is happening, the input impedance of the amplifier drops as does the maximum impedance realizable in the plate circuit.

Numerous factors contributing to the reduction of amplifier output at vhf and uhf can be listed and divided roughly into three groups:

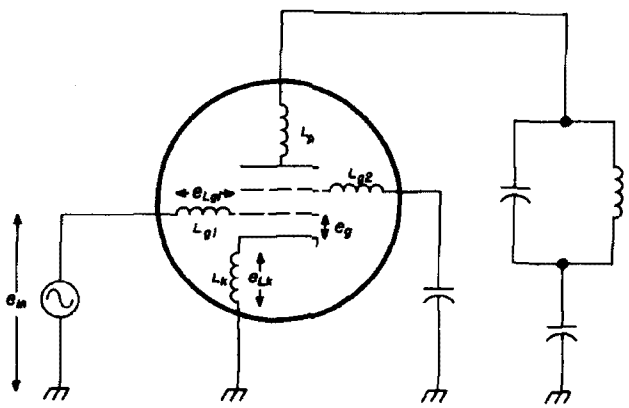
1. Circuit-reactance limitations
2. Circuit and tube loss limitations
3. Electron transit-time limitations.

circuit-reactance limitations

At the very-high and ultra-high frequencies there exists a situation which is quite different from that which exists at low frequencies. At low frequencies the electrical circuits and the tube are quite distinct. As frequency increases, this ceases to be true, and it is found that part of the resonant circuit exists within the tube (**fig. 1**). The electrode leads, while they are normally short and have large surface area in modern tubes, have a small but finite inductance. As the frequency of operation is raised, the reactance of the **lead inductance** will become quite appreciable, often reaching undesirable proportions in the vhf-uhf regions. The effect of lead inductance is to create a voltage drop such that the applied voltage across the external terminals of the tube will not entirely appear across the electrodes. Driving voltage is thus lost across the lead inductance.

In addition, while the interelectrode capacitances may be small, at ultra-high frequencies they approach a large fraction of the capacitance required to establish resonance in an external circuit. As such, the interelectrode capacitances represent a limitation in terms of actual operation as the external tank circuit may "disappear" within the tube. The combination of the electrode-lead-inductance and the interelectrode capacitance may cause an **internal resonance** in the uhf region.

fig. 1. **The total input voltage does not act upon the grid-to-cathode electron stream because of grid and cathode lead inductance—the voltage across them subtracts from the input voltage.**



A typical internal resonance experienced in a large power tube is the circuit consisting of the control-grid cage and mounting cone, and the screen-grid and cone. These two assemblies can form a quarter wavelength long tuned-line circuit shorted at the tube envelope by the capacitor consisting of the two contact rings and the dielectric material used in the envelope. The internal resonant frequency could be in the range of 1400 MHz for a five- to ten-kilowatt tube. This could lead to a 1400-MHz parasitic oscillation.

The smaller the tube, the higher the resonant frequency. All tubes will have internal resonances and the designer must move them out of the normal usable frequency range or load the circuit in such a way as not to degrade the performance within the rated frequency range. Even if resonances do not occur, the combination of reactances within the tube may constitute an undesired network that creates a mismatch between the driver and the tube electrodes.

magnitude of lead inductance

The most important reactance encountered in a vacuum tube is that associated with the lead inductance. An estimate of the magnitude of lead inductance may be made using the following equation:

$$L = 1.00508\ell \left(2.303 \log_{10} \frac{4\ell}{d} - 1 + \frac{d}{2\ell}\right)$$

where L = inductance (microhenries)
ℓ = wire length (inches)
d = wire diameter (inches)

This equation assumes no mutual inductance with some other nearby lead or wire.

As an example of how great lead inductance can be, consider the case of a lead that is 0.1-inch in diameter and one-inch long. This lead will have an inductance of 0.015 μH and an inductive reactance of 47 ohms at 500 MHz. An inductive reactance of this magnitude, for example, between the screen element and the screen by-pass capacitor outside the tube can cause stability problems. A voltage drop exists across the

screen reactance caused by the current flowing through it which is the vhf current charging the output capacitance of the tube. This voltage is impressed upon the screen, which thereby is removed from ground potential, disturbing the grid-plate isolation normally afforded by this element.

This source of positive input resistance can be better understood by realizing that it results from a signal appearing across the cathode lead inductance, driving the cathode a small amount, as in a cathode-driven amplifier. (In a cathode-driven amplifier, the alternating current component of the plate

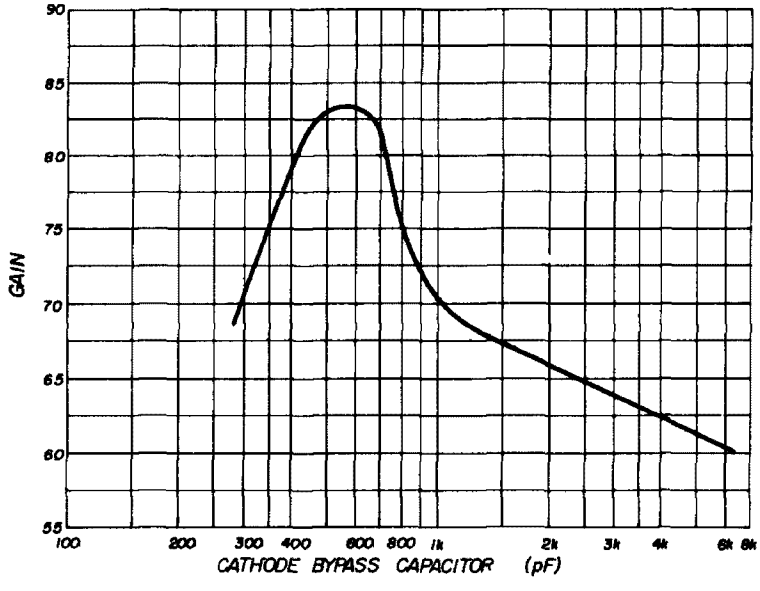


fig. 2. Gain of a 5CX1500A 100-MHz class-B amplifier with different values of cathode bypass capacitance. Gain peak occurs when bypass capacitor is series resonant with tube, socket and cathode lead inductance.

cathode lead inductance and input loading

Tube gain will also be adversely affected due to a reduction in input resistance because of the cathode-lead inductance. A small amount of cathode-lead inductance in conjunction with the grid-to-cathode capacitance of a vacuum tube will cause a resistive load to appear across the input of the tube. The magnitude of this added shunt input resistance is:

$$R \cong \frac{1}{\omega^2 L_k C_{gk} g_m}$$

R = input resistance due to cathode-lead inductance
$\omega = 2\pi f$ where f is frequency (hertz)
L_k = cathode lead inductance (henries)
C_{gk} = Grid-to-cathode capacitance (farads)
g_m = transconductance (mhos)

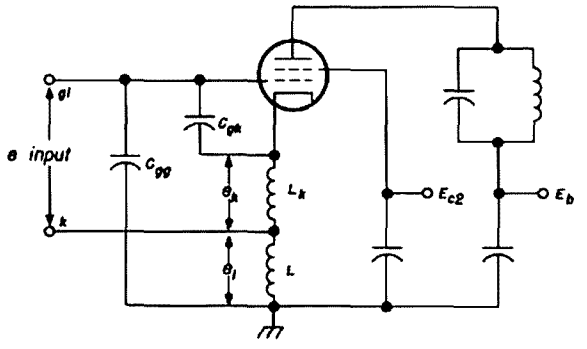
current has to flow through the cathode lead inductance to reach the cathode. Since the driving signal is in series with the output load through the cathode-to-plate resistance of the tube, some of the power in the load is supplied by the driver.) Unwanted feedthrough power is thus supplied by cathode lead inductance in a grid-driven stage.

The above equation shows that as the frequency is increased, the grid-loading due to cathode-lead inductance increases. For example, if the frequency of the amplifier is tripled from 144 to 432 MHz, the input resistance of a particular tube at 432 MHz would be one-ninth the resistance at 144 MHz. If it is desired to drive the 432 MHz tube to the same maximum grid voltage swing, then nine times the power is required just to cover the input loading due to cathode-lead inductance.

There will be other losses such as skin

effect, dielectric, radiation, transit time and the circuit reactance effects shown in **fig. 1.** The amount of grid loading due to cathode-lead inductance may be reduced somewhat by separating the input and output circuit paths back to the cathode. Some miniature tubes and low power transmitting tubes, such as the 4CX250B, 6146 and others, have multiple cathode leads to minimize the cathode-

fig. 3. Cathode lead neutralization.



lead inductance effects. Transistors also have this problem, only it is called emitter-lead inductance instead of cathode-lead inductance.

vhf/uhf cathode-lead inductance neutralization techniques

It is possible to neutralize the effects of the cathode-lead inductance by choosing the value of cathode by-pass capacitor so it will be **approximately** series resonant with the total lead inductance (tube, socket and circuit inductance). This technique is particularly effective in low noise stages of vhf/uhf receivers.

Fig. 2 illustrates stage gain as a function of cathode by-pass capacitance value in a high power vhf amplifier. This particular graph was obtained from an experimental 100-MHz, 5CX1500A amplifier running in class B. The 5CX1500A cathode lead, socket and circuit lead inductance was measured to be just over four nanohenries. A capacitance of 637 picofarads was calculated to be necessary for cathode lead inductance neutralization. The graph agrees fairly well with the calculated data.

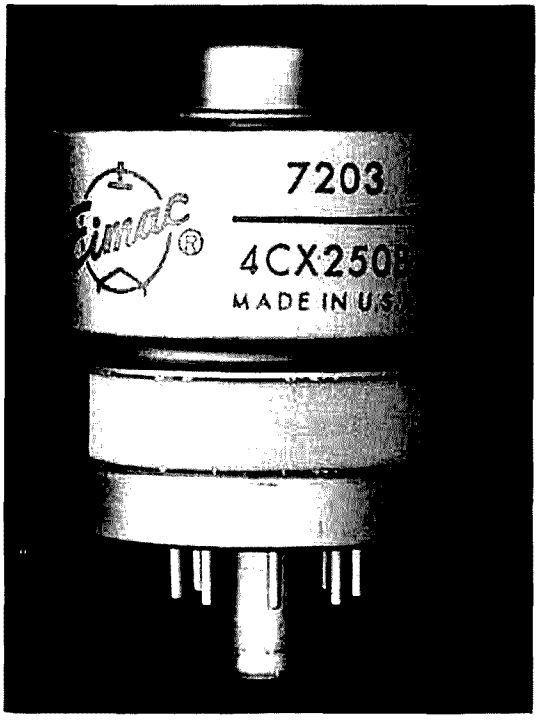
A neutralization technique described in the October 1939 issue of **Electronics** is of interest (**fig. 3**). I have had no personal experience with this technique, but it does appear to have merit. The voltage drop e_L across L, caused by the cathode current, is reversed in polarity with respect to e_k in the sense that C_{gg} and L are in series between grid 1 and the cathode. Thus, the current flowing back to the grid through C_{gg} is 180 degrees out of phase with the applied voltage, e_{input}. For a certain value of L the currents through C_{gk} and C_{gg} will be equal as well as opposite in phase; thus, the conductance between grid 1 and the cathode is zero. For neutralization of the cathode lead inductance the following ratio must be met:

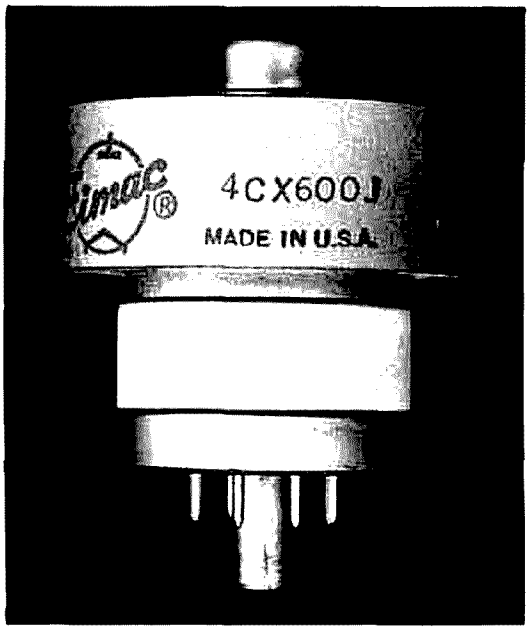
$$\frac{L}{L_k} = \frac{C_{gk}}{C_{gg}}$$

For the 5CX1500A 100-MHz amplifier component values for neutralization of the cathode lead inductance could be:

$L_k = 4$ nanohenries
$C_{gk} = 35.8$ picofarads
$C_{gg} = 10$ picofarads
$L = 14$ nanohenries

The well known 4CX250B—rated to 500 MHz.





This tube has twice the cathode area of the 4CX250B and provides better than —40 dB intermodulation distortion.

L_k and C_{gk} are typical characteristics of the 5CX1500A and SK840 socket while C_{gg} and L are added components.

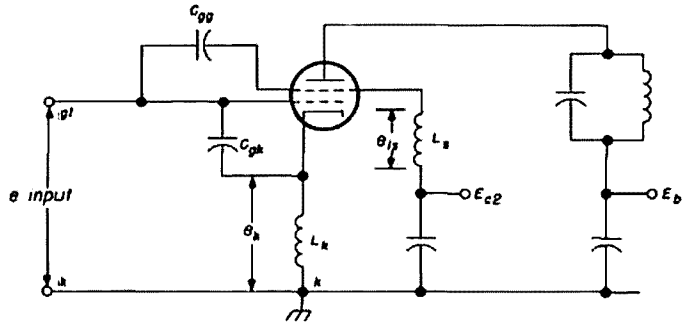
A very similar method of neutralizing that part of input resistance caused by the cathode lead inductance uses an inductance in the screen-grid lead and has been often used in 6146-type gear at 144 MHz (**fig. 4**). Basically, this circuit is the same as **fig. 3**. There are two differences: the point K (cathode pin) is now at ground potential and only screen current flows through the inductance L_g.

Other circuits may be used to minimize the effects of the cathode-lead inductance. The grounded-grid, cathode-driven (or probably more correctly called the "grid-separation" circuit), is often used (**fig. 5**). In this case the cathode is driven while the grid in a triode, or the grid and screen grid in a tetrode, are operated at some low rf potential. The grid structures then act like a shield between the input and output circuits. The main advantage, as far as cathode-lead inductance is concerned, is that this inductance is now just another inductor in series with, and therefore a part of, the input tuned circuit. There are other advantages that can be credited to the "grid separation" amplifier which will be discussed in the section on interelectrode capacitances.

intermodulation distortion and input loading

An appreciable amount of input loading can increase intermodulation distortion. Any tube plate-characteristic non-linearity will cause a variation in this input loading with signal level and thereby present a varying load to the driver, thus causing increased distortion in the drive voltage.

fig. 4. Neutralizing cathode lead inductance with an inductor in the screen lead on the tube side of the screen bypass capacitor.



screen lead inductance

The screen lead inductance between the screen element and the screen by-pass capacitor may help or hinder the operation of an amplifier. Below the **self-neutralizing frequency**[1,2] of the tube, the screen lead inductance is usually detrimental to the stability of the amplifier, as the rf current flowing through this inductance will cause an unwanted rf voltage to be developed. The point where the screen bypass is connected to the screen terminal may very well be at rf ground potential, but the potential of the screen itself may be varying above and below ground by the magnitude of voltage developed across the screen lead inductance, eL_g (**fig. 4**).

The magnitude of the developed voltage depends on the inductance of the screen lead and the frequency of operation. The higher the frequency, the greater the inductive reactance and the greater the rf current through this inductor. The current is greater

because the capacitive reactance of the output capacitance of the tube will be smaller, and the capacitor will be charging to the same rf plate voltage swing each cycle.

At operating frequencies below the self-neutralizing frequency of a tetrode, screen inductance may be added in order to neutralize the amplifier. That is, the self-neutralizing frequency is lowered to the operating frequency.[3] At operating frequencies above the self-neutralizing frequency of the tube, a series capacitor is sometimes added to move the self-neutralizing frequency up to the operating frequency (**fig. 6**). As was briefly touched upon previously, sometimes it is desirable to add a certain amount of screen lead inductance to neutralize the cathode-lead inductance portion of the input resistance.

control-grid lead inductance

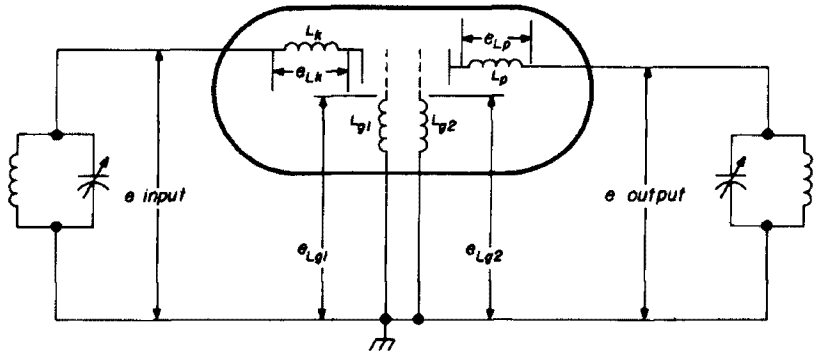
The **control-grid lead inductance** in a grid-driven amplifier is usually not of much concern as the relative magnitude of the tube lead inductance as compared to the external inductance added to it to attain resonance is very small. The control-grid lead inductance is wholly a part of the input resonant circuit with no current being induced from the output circuit, and it becomes part of the input tuned circuit. An exception to this is in the case of a vhf/uhf **distributed amplifier.** In this application an active filter is designed using all of the tube lead inductances and interelectrode capacitances. The control-grid inductance is important in this case, and modern distributed amplifier tubes are made with four grid leads available to the equipment designer. He then has the choice of the number of leads to use to provide the inductance he needs for his design.

In a cathode-driven (grounded-grid) amplifier, control-grid inductance is very important. Just as in the case of the screen-lead inductance in a grid-driven tetrode amplifier, the control-grid inductance in a cathode-driven amplifier may aid or hinder the designer. The control-grid inductance may cause instability, a loss in drive voltage due to the voltage divider effect (**fig. 1**) or it may be used to provide a method of neutralizing the amplifier.[2]

plate lead inductance

The plate in modern tubes used for vhf/uhf operation is usually designed with a massive anode structure in order to dissipate the heat that is generated in this element of the tube. For this reason, plate lead inductance is usually low enough so it is not of any great concern. If normal good engineering practice is followed in designing vhf/uhf circuitry, the plate lead inductance becomes inconsequential.

fig. 5. Cathode-driven or grid-separation amplifier.



interelectrode capacitance

In addition to lead inductance, interelectrode capacitance plays an important role in the operation of tubes in the vhf/uhf regions. Interelectrode capacitances due to active parts of the tube structure are incapable of reduction beyond a certain point. However, in many tubes the interelectrode capacitance results largely from capacitance between leads in areas of the tube where electrons do not flow. It is the job of the tube designer to reduce this unnecessary capacitance to a minimum.

input capacitance

The input capacitance of a grid-driven tube is the sum of the grid-to-cathode and grid-to-screen capacitances. The larger the input capacitance, the greater the drive power must be. This can be explained by the very large increase in input charging current necessary to charge the input capacitance. As the frequency increases the reactance of the input circuit becomes smaller, and for the same peak grid voltage, the charging current must increase. As power is proportional to the square of the current, the power lost in the input circuit necessarily increases as the charging current rises. The driver must supply this extra power.

To reduce this loss, the circuit designer must keep the input capacitance down and may limit the magnitude of the peak grid voltage. Peak grid voltage can also be minimized by operating with less bias. Quite often in certain amplifiers the class-B mode is more desirable than class-C operation. Reducing the peak grid voltage and the charging current will reduce the amount of power that must be dissipated by the control grid. Radio frequency power dissipated by the control grid unfortunately cannot be measured by the dc meters on the front panel of the amplifier, so the operator has no means of knowing if charging currents cause excessive grid temperature. High input capacitance also limits the bandwidth of the input circuit. For those applications requiring large instantaneous bandwidth, great care must be taken in the design of the equipment.

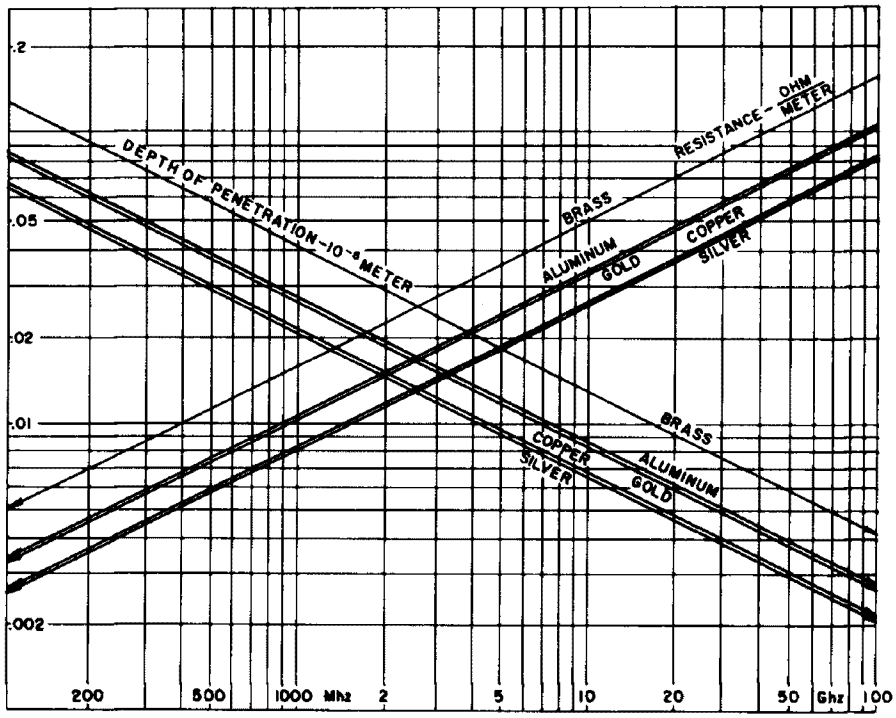
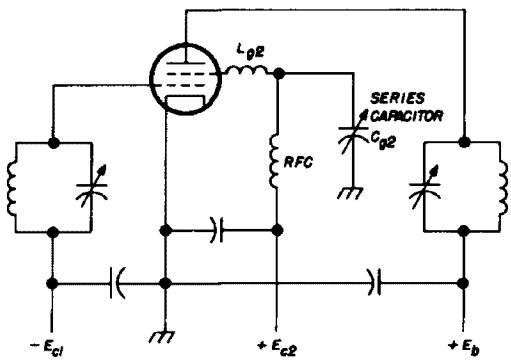


fig. 7. Chart for computing resistivity and depth of penetration for metallic conductors between 100 MHz and 100 GHz.

fig. 6. Screen neutralization at operating frequencies above self-neutralizing frequency of tube.



The grid-separation or cathode-driven amplifier offers quite an advantage as far as input capacitance is concerned. The input capacitance consists only of the cathode-to-grid capacitance. For the same tube in the cathode-driven configuration, the input capacitance will be roughly half that value in the grid-driven circuit. This is quite an advantage for applications requiring wide bandwidths.

output capacitance

The output capacitance of a power tube is an important factor in determining what plate-load resistance can be used. This in turn determines the stage gain and power output that is available. The equivalent shunt resistance (plate-load) of a parallel resonant circuit can be written as $R_L = Q/2\pi fC$ or $R_L = 2\pi fLQ$ where R_L is the plate-load resistance, Q is the loaded Q of the resonant circuit and f is the resonant frequency. For

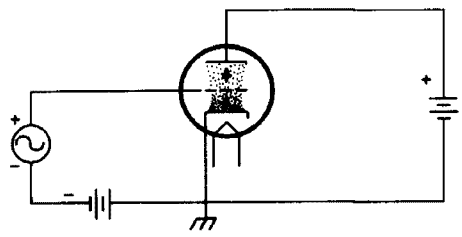


fig. 8. The effect of transit time on grid losses. At the instant shown, the grid potential is increasing in a positive direction and there is a consequent disproportionate number of electrons between grid and cathode so there is electron flow from the grid even though it may be negatively biased.

operation at a given frequency, to **increase** the shunt resistance it is necessary to **decrease** the shunt capacitance. This can be done to a point by reducing the circuit capacitance and increasing the tank coil inductance, maintaining the same frequency of resonance. Eventually, this process is limited by the fact that the capacitance external to the tube has been reduced to zero; the shunt resistance is finally determined by the tube interelectrode capacitance. The larger the interelectrode capacitance, the smaller the shunt resistance that can be realized. Accordingly, power output tends to drop off as the load resistance, or as the square of frequency, as frequency increases.

There are other problems brought about by the effect of output capacitance. The output capacitance must be charged and discharged during each cycle of the radio frequency. Again, as the frequency increases, the reactance of the output capacitance decreases. Therefore, with the same value of peak rf plate voltage, the current flowing through the output capacitance must increase as the frequency increases. The output capacitance of a tetrode is made up of the screen grid structure, the plate structure and the tube envelope.

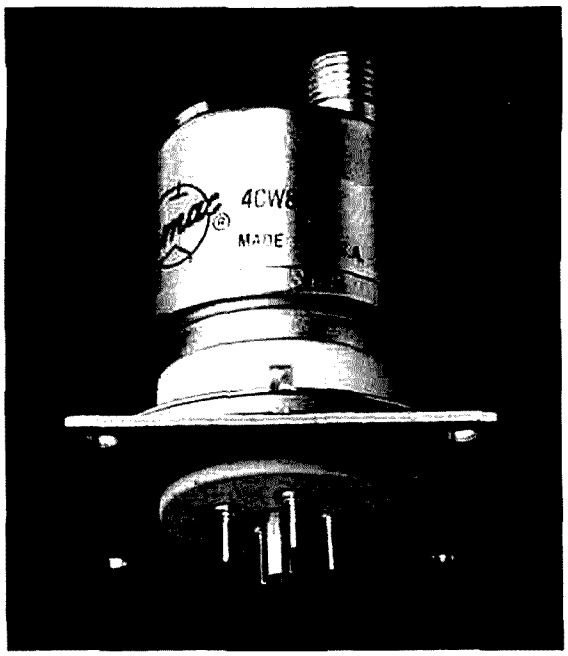
The charging currents must flow over the surface of these components of the tube, all of which have varying degrees of rf resistance. It is possible for the charging currents to exceed the dissipation rating on the screen grid even though the dc meters indicate all is within ratings. It is advisable in vhf/uhf circuits, therefore, to try to achieve the lowest usable value of peak rf plate voltage by using the lowest plate load resistance and drawing the highest plate current consistent with desired output power and efficiency. Running the tube in this manner will lighten the rf stress on the tube seals and reduce the rf current in the screen grid structure thus providing for a potentially longer tube life.

feedback capacitance

The feedback capacitance in a grid-driven amplifier is the capacitance from anode to the control grid. The higher the frequency of operation, the greater the chance for instability due to rf feedback from the output circuit through feedback capacitance to the input circuit. In the vhf/uhf region this capacitance and other tube capacitances and inductance must be adjusted to provide for neutralization by added circuit components.[4]

The grid-separation amplifier helps minimize the effects of the feedback capacitance. The feedback capacitance in this configuration is very much less than the grid-driven case since it is the capacitance from anode to cathode with the grid, or grids, shielding the output from the input. In some applications no neutralization will be required. Other applications may require quite extensive neutralization.[3,5]





Distributed-amplifier tube discussed in text. By choosing the number of control-grid leads, the designer can control lead inductance. The tube in the right-hand photo has a screen by-pass capacitor installed. Center pin is one of the heater pins; the square tab on the capacitor is the other heater connection and the cathode. The threaded pins *are* control-grid leads.



Tube designed for large phased-array radar system. Two of these tubes were used in the 1000-watt 432-MHz amplifier used by WA6LET with a 150-foot dish for moonbounce contacts in 1965.

circuit and tube-loss limitations

The power losses associated with a tube and circuit all tend to increase with frequency. In the vhf/uhf region, almost all radio-frequency current flows in the surface layers of a conductor because of skin effect (**fig. 7**). The resistance and rf losses in a conductor increase with the square root of frequency because the layer in which the current flows decreases in thickness as the frequency increases. Insulating supports in the tube and external circuit have losses associated with the molecular movements produced by the electric fields. These dielectric losses will usually vary directly with frequency. Also, there will be additional losses due to the *radiation of energy from the wires and leads* carrying rf current. The power radiated from a short length of wire carrying current increases as the square of the frequency.

All these factors contribute to a general reduction in tube and circuit efficiency as operating frequency is increased. In the manufacture of power tubes the resistance losses are reduced by increasing conductor surface area and by proper choice of lead materials. Dielectric losses are reduced by

selection of envelope and insulating materials. Support insulators are positioned, when possible, out of high voltage fields. Radiation losses are reduced by constructing vhf and uhf tubes and circuits so as to be totally shielded. At times it is prudent to use concentric line construction techniques so that tube and circuit fields are entirely confined.

transit-time limitations

Electron transit-time effects can contribute to reduced tube output in many ways. Transit time is the finite time an electron takes in going from the cathode to the grid. If the transit time (a function of grid-to-cathode distance and grid-to-cathode voltage) is an appreciable fraction of one ultra-high-frequency cycle, then an electron in transit in the grid-to-cathode region might be still heading for what was once a more positive location than the cathode surface it just left, but now finds the grid may be less positive or perhaps even negative (**fig. 8**). As a result of transit-time effect in the cathode-grid region, there will be a dispersal of "out of step" electrons. Because of this dispersion of the electron stream, the plate-current pulses are not as sharp as the current pulses liberated from the cathode.

In addition, energy is required to accelerate the electron towards the anode, and this energy is supplied by the driver. As the operating frequency is raised, more energy is required because the grid-input resistance due to transit time varies inversely as the square of the frequency. That is, if the frequency is doubled, the input resistance due to transit time effects will be one-fourth that at the lower frequency. The extra power required to overcome transit-time loss due to grid-input resistance is supplied by the driver and appears as lost drive-power—required but put to no practical use other than to heat the tube seals and waste precious exciter output.

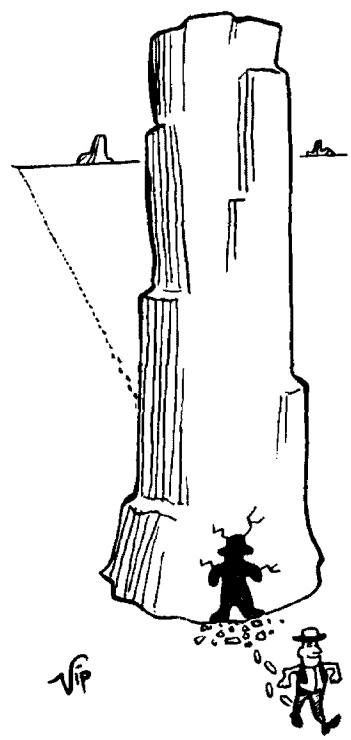
Paradoxically, transit-time loss and cathode-lead inductance loss (both of which cause input loading) are not all evil because they often tend to stabilize a "wild" stage. Cure of the trouble may lead to higher stage gain with the possibility of oscillation and instability!

Circuit loading can be used to achieve stability and this is often done at the lower frequencies by adding resistors across i-f transformers, for example. The circuit designer, therefore, finds that a tradeoff exists between loading and stage gain that will work to his advantage or disadvantage, depending upon his ability to analyze the circuit. There is often more than one successful path to a proper design and the good circuit engineer has several alternative paths (in his head, if not on paper) to choose from. A candid realization of uhf/vhf effects will help the circuit designer obtain maximum power, efficiency and reliability from his equipment.

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ham radio



High-power linearity with Eimac power-grid tubes.

solid-state circuits for single sideband

More and more transistor circuits are creeping into commercial ssb gear—here are some of the circuits that are currently being used

Forest H. Belt, 2610 Whittier Avenue, Louisville, Kentucky 40205

Transistors haven't taken over ham equipment as quickly or as thoroughly as they have some electronic gear, but semiconductors of all kinds are steadily working their way in. You see more solid-state construction in homebrew rigs than in commercial ones. Nevertheless, when you look over the range of ssb equipment now available, you find there's hardly a circuit that hasn't gone solid-state in some brand or some model. No serious ham can help being interested.

At first, only small size and portability (battery operation) were the reasons for using transistors. But semiconductors have been around for 20 years now, and they are more reliable and stable than ever before. With few exceptions (high power, extremely high frequency) transistors can now perform most tube functions—and in many cases do a better job. Furthermore, the field-effect transistor (FET) has made high-impedance solid-state circuits possible.

Here are some places you'll see transistors in single-sideband transmitters and receivers.

speech or af amplifier

The two-stage amplifier of **fig. 1** is typical of audio transistors used in sideband rigs. This circuit is from the Galaxy V Mark 2 transceiver. Q9 and Q10 are speech amplifi-

ers between the microphone and the balanced modulator. Power and frequency demands are modest, and hundreds of inexpensive transistor types can do this job.

The dc supply hookup is simple—positive voltage to the collectors, and emitters returned to ground through stabilizing resistors R105 and R109. R106 and R112 drop the supply voltage to the right collector potential, and C157 and C124 decouple the two stages from each other and from the rest of the transceiver.

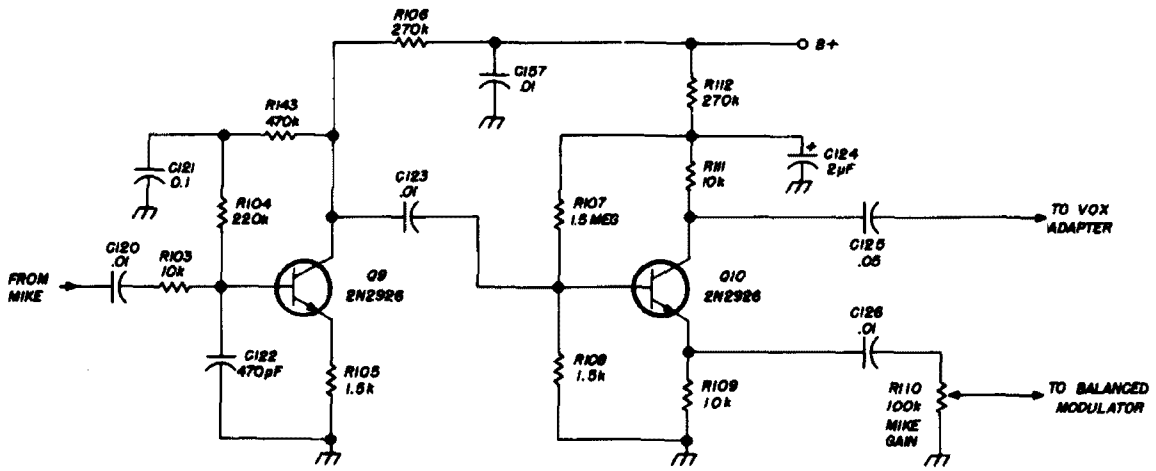
Q9 gets base bias through R143 and R104, with C121 for decoupling. Q10 uses R107 to carry base bias; R107 is a voltage divider with R108.

audio output stage

Audio power requirements aren't stringent in ssb receivers; a watt or two is usually sufficient to drive the speaker. Many sets use a pair of transistors in class-B push-pull service—like the one in **fig. 2,** from the Gonset 910A transceiver.

Since these transistors are pnp types, they require negative collector voltage (with respect to emitter). In this hybrid rig, however, supply voltage is positive with respect to chassis. The collectors of Q10 and Q11 are connected to ground through the low dc resistance of T5's split primary. The emitters go to the 12-volt line through R32, a stabilizing resistor. The collectors are thus negative

fig. 1. Microphone amplifier.



The audio signal is brought from the microphone through blocking capacitor C120 and is applied to the base through a filter consisting of R103 and C122. This filter removes any stray rf, but has little effect on speech signal. Emitter resistor R105 is unbypassed; that adds a little degeneration to improve linearity in the stage. The audio output from Q9 develops across collector load R106.

This output signal is fed to the base of Q10 through C123. Q10 is a phase splitter, with collector and emitter resistors (R111 and R109) equal. The collector signal goes to a VOX adapter. The emitter signal is fed through C126 and microphone gain control R110 to the balanced modulator.

with respect to the emitters. Base bias comes from a voltage divider (R30-R31).

The signal from the audio driver stage is coupled to Q10 and Q11 by the split secondary windings of input transformer T4. The collectors feed their output signals to the split primary windings of T5. The secondary matches a 3.2-ohm speaker.

crystal bipolar oscillator

Raytheon's Sideband Engineers SB-34 transceiver uses a germanium bipolar pnp transistor as a 456.38-kHz oscillator. On transmit, it generates the carrier frequency, and on receive it supplies the carrier-reinsertion signal. Output from this stage is fed to the balanced modulator, which is also the bal-

anced detector for receiving.

The circuit—shown in **fig. 3**—is a Pierce crystal oscillator. Feedback is from collector to base via the crystal.

As in the preceding circuit, the supply voltage is connected to the stage in what some people call an upside-down arrangement, which is used a lot with pnp transistors. The emitter and base are positive with respect to chassis, and the collector is grounded (through L11). The collector is thus negative with respect to the emitter.

Base bias is taken from divider R18-R19. The emitter stabilizing resistor (R20) is bypassed by C19, to prevent degeneration. Output voltage is developed by the collector across L11. Since the output is untuned, Q12 is operated class A, with a fairly clean sine-wave output.

fet crystal oscillator

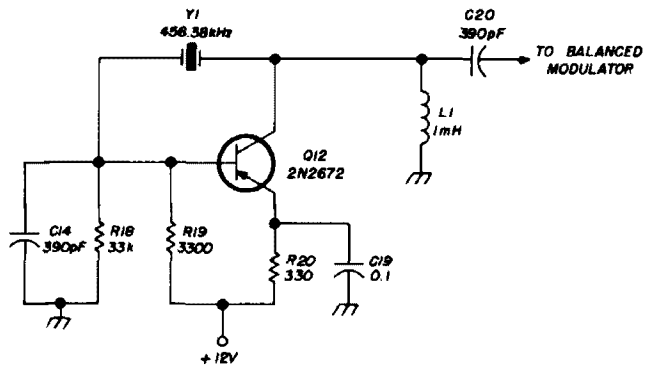
One advantage of the field-effect transistor (over the bipolar) is that the FET's high impedance loads a crystal very lightly, making a more stable oscillator. FETs also have better temperature stability than bipolars.

These advantages are put to good use in the Drake R-4B receiver, which uses an n-channel junction FET as a calibration oscillator. The circuit is shown in **fig. 4.** It's another Pierce type, with feedback through the crystal from drain to gate. C119 is a trimmer to allow zero-beating the 100-kHz oscillator with WWV.

R73 returns the gate to ground for dc. Source bias is used, and source resistor R72 is bypassed for rf by C120. Output voltage is developed across drain load R155, decoupled by C189. C188 feeds this rf to a buffer amplifier. C138 lowers the drain-circuit impedance, thereby stabilizing the oscillator. The output is very high, so the shunting effect of C138 does no harm.

Q8 is operated class A and produces a clean sine wave at 100 kHz. Other stages divide this down to 50 and 25 kHz. The last stage produces lots of harmonics, putting harmonics every 25 kHz across the bands of interest. By rearranging jumpers in the divider stages, the user can get 25-, 50-, or 100-kHz markers to calibrate from.

fig. 3. Pierce type crystal oscillator.

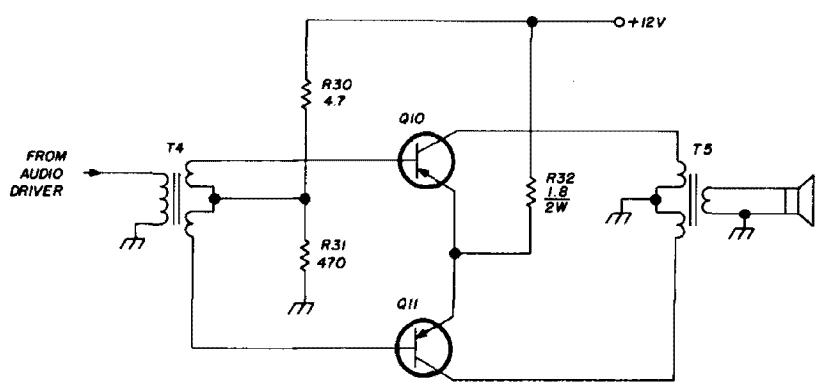


voltage-tuned vfo

In the past, vfo's have been LC-tuned, usually with an air-dielectric capacitor as the variable element. Some new transistor vfo circuits, however, use a voltage-variable capacitor (also called a varactor or varicap). The Gonset 910A transceiver uses a varactor to tune a vfo over the range of 5.5-6.5 MHz.

The circuit is shown in **fig. 5A,** and it's somewhat complex. Look at how it's redrawn in **fig. 5B,** though, and you'll recognize a

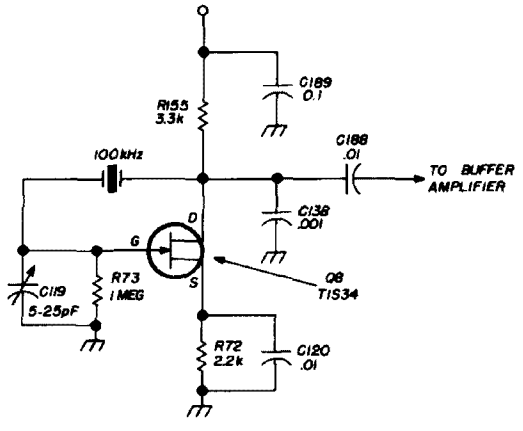
fig. 2. Class-B audio output stage.



Colpitts oscillator; C32 and C34 form the feedback divider. L1 and all those capacitors are in series across the divider.

Q17's collector is negative with respect to its emitter, as befits a pnp bipolar. R33 is the load resistor, connected from collector to ground. R46 is the emitter stabilizing resistor, and connects the emitter to the positive 10-volt line. Base bias comes from a divider

fig. 4. Pierce oscillator using a field-effect transistor.



(R45-R34) between the 10-volt line and ground.

Still referring to **fig. 5B,** note that in the tank circuit there is only one variable element. This symbol represents six capacitors and a variable-capacitance diode. In **fig. 5A** you can see those parallel capacitors on the left side of the diagram. C26, C27, and C28 furnish the bulk of capacitance that resonates with L1. C29 and C30 are trimmers for aligning the vfo. C31 is a blocking capacitor, so the control voltage can be connected to D8 without affecting the base of Q17.

A voltage divider (R68, R47, and the vfo tuning control), connected from the 10-volt line to ground, supplies the control (tuning) voltage. As the slider of the tuning control is moved, it varies the bias applied to the diode. As the capacitance of D8 varies, it changes the frequency of the oscillator over a 1-MHz range. L2 and C98 decouple the tank circuit from the control-voltage line.

i-f amplifier

For an example of a simple i-f amplifier, look at **fig. 6**—the circuit used in Gonset's 910A transceiver; frequency is 9 MHz.

You should recognize the upside-down dc supply circuit used for Q5. Its collector is grounded for dc through T2's primary, and the emitter is returned to the 12-volt positive line through emitter stabilizer R16 (bypassed for rf by C12) and the rf gain control. Base bias is applied through R58, but is overridden during certain signal conditions by voltage from the agc line. Stage gain is varied two ways: by the manual rf gain control and by the agc line.

Transistor i-f transformers—such as T10 and T2—often use a lot of turns on the tuned winding (the primary in this case) to improve the Q of the tuned circuit. With many turns on the coil, the parallel capacitor can be small. Also, the slug-type tuning is more effective. But, because bipolar transistors are low-impedance devices, the primary of T2 is tapped at a low-impedance point for the collector connection. The low-impedance base of the next stage is fed from a low-impedance, few-turn secondary.

two-way i-f amplifier

The Raytheon SB-34 transceiver uses a novel i-f amplifier stage at 456 kHz. During transmission, the stage amplifies the dual-sideband signal from the balanced modulator and applies it to the mechanical sideband filter. During reception, the direction is reversed. Incoming rf is heterodyned down to the 456-kHz i-f and passed through the mechanical filter. The i-f stage amplifies these signals and passes them to the balanced modulator, which then functions in reverse as a product detector.

The circuit of this bilateral i-f amplifier is shown in **fig. 7.** The emitters of the two transistors share a common stabilizing resistor (R33) and by-pass capacitor (C29), and both emitters are tied to the 12-volt positive line. Q5's collector is returned to dc ground through the mechanical filter (not shown). Q6's collector is returned to dc ground through the secondary of T2.

As you may have guessed, one transistor is on during transmit, and the other is off. They switch during receive. The swapping is done by changing base bias. In the transmit

mode, line A is grounded. R31 and R32 become a voltage divider from 12 volts to ground, forward-biasing Q5 and turning it on. The sidebands of 456 kHz are fed from the balanced modulator through T2, and then through C28 and R30 to Q5, which amplifies them and sends them on to the mechanical filter.

Line B meanwhile is 12 volts positive. Divider R37-R38 has 12 volts at both ends. Q6 is therefore not forward-biased, and does not conduct.

In the receive mode, line A gets 12 volts, which turns off Q5. Line B is grounded, so now divider R37-R38 develops bias at the base of Q6. Rf coming from the mechanical filter goes through C30 to the base of Q6, where it is amplified and fed through T2 to the balanced-modulator/product-detector.

rf amplifier

In one commercial transceiver, a simple rf amplifier is used. Its circuit is shown in **fig. 8.**

The dc circuit is pretty much what you've seen before. No emitter resistor is used; the emitter is tied directly to 12 volts—although it is bypassed for rf by C106.

Base bias is developed by divider resistors R97 and R98; but notice: they're connected between the 12-volt line and the agc line. This keeps the rf stage from overload by applying agc action.

Rf from the antenna is fed through the T/R relay to tuned circuit L17-C129, which is broadly resonant across the transceiver's operating range of 49.975 to 54.025 MHz. The rf input is tapped down on L17 to match the 50-ohm antenna.

Another tap on L17, a little higher, matches the base impedance of Q21. From there, dc blocking capacitor C107 feeds the rf to Q21.

A tap on L16 matches the collector impedance of Q21. The resonance of tank circuit L16-C105 is broadened by R95. The rf output signal developed across the tank is fed through C104 to the first mixer.

two-way mixer

Earlier, you saw the two-way i-f amplifier used in the Raytheon SB-34 transceiver. The same principle is used in that unit's vfo mixer. The circuit is shown in **fig. 9.**

The dc hookup is about the same as in the i-f amplifier in **fig. 7.** The collectors are returned to ground through T3 and T6, and the transistors share a common emitter resistor (R45) and bypass (C43). Base bias is developed by dividers from the 12-volt line to control lines A and B.

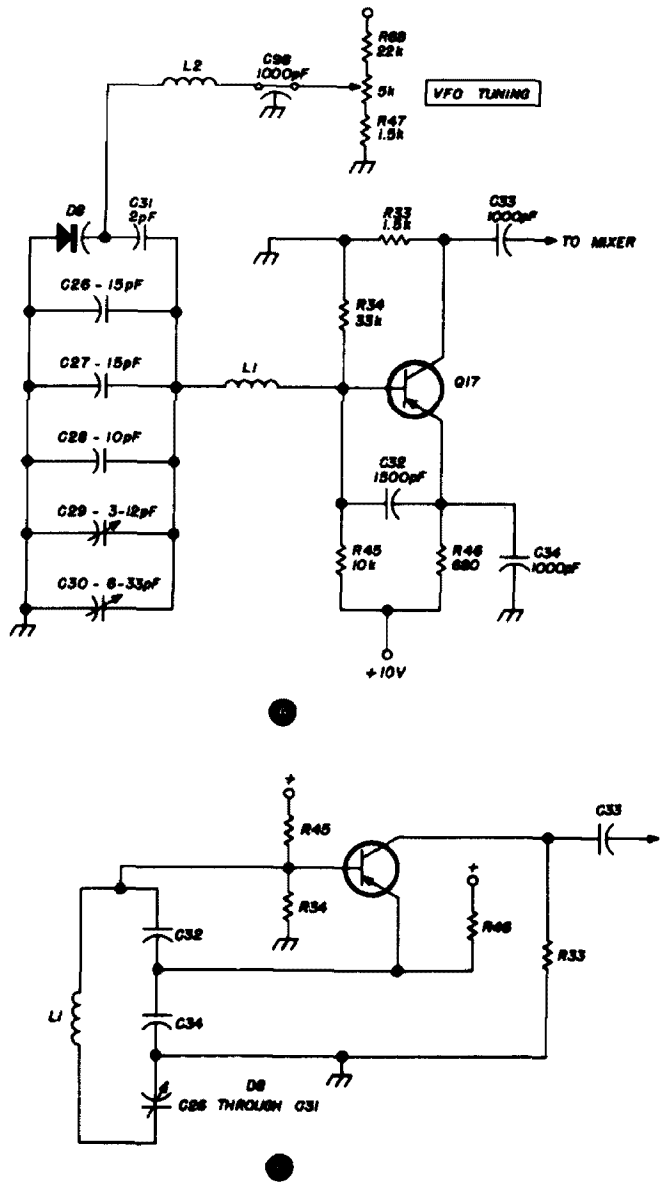
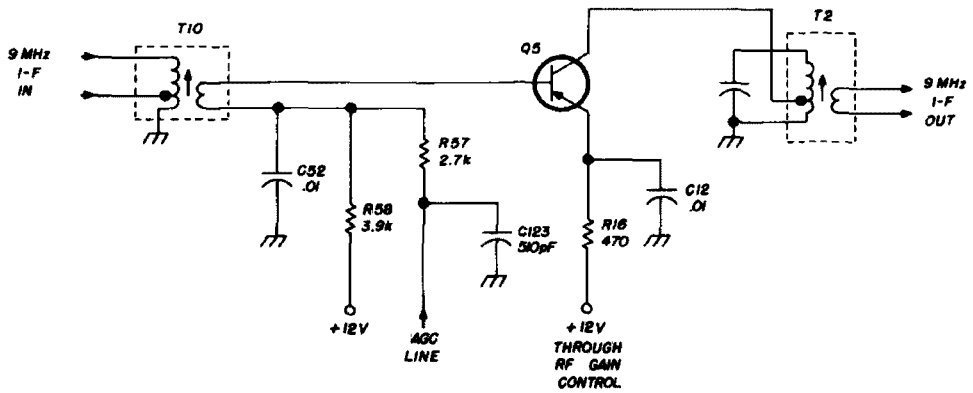


fig. 5. Voltage-tuned vfo; simpler version is shown in B.

In the transmit mode, Q7 is on; it amplifies the sideband signal, sending it on to the hf mixer and ultimately to the power amplifier and antenna. During the receive mode, Q8 is on; it amplifies incoming rf and sends it to the mechanical filter.

But the purpose of this stage is to add the vfo output to the i-f signal. The vfo operates between 5457 and 5707 kHz. Transformer L4 the vfo signal, producing a difference frequency of 2289 kHz, which is fed to the next mixer stage.

fig. 6. Simple intermediate-frequency amplifier.



couples its output to the emitter-base circuits of both Q7 and Q8.

For transmitting, the sideband signal passes from the mechanical filter through T3 to Q7. There, it is heterodyned with the vfo (5457-5707 kHz). The new difference frequency is between 3175 and 3425 kHz, and this signal is applied to T6. Of course, T6 accepts only the specific difference frequency produced by the vfo. It's tuned by the A and B sections of C48, which are on the same shaft as the vfo's tuning capacitor.

For receiving, the system works in reverse. The incoming rf is in the range of 3175 to 3425 kHz. It's tuned by C48 and T6 and passed to Q8. There, it is heterodyned with

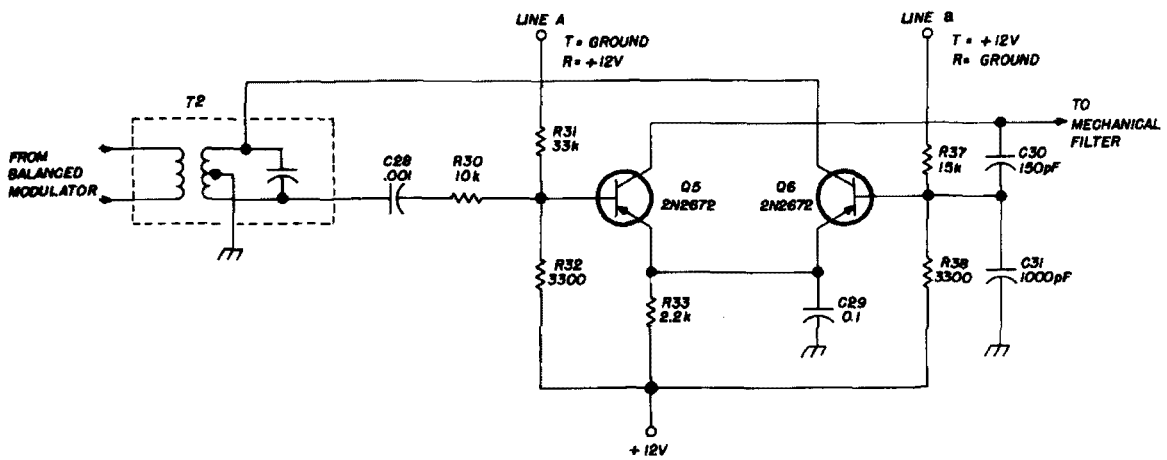
product detector

In the Gonset 910A, a transistor is the product detector for ssb (a diode is the a-m detector). The emitter-follower circuit is shown in **fig. 10.**

The rf sidebands around 9 MHz are brought through T3, whose secondary is a low-impedance link that matches the low base impedence of Q7. C16 keeps dc out of T3 while coupling the rf to Q7's base. A steady signal from the 9-MHz crystal oscillator is also fed to the base of Q7, where it heterodynes with the sideband signal from transformer T3.

The collector of Q7 returns to ground

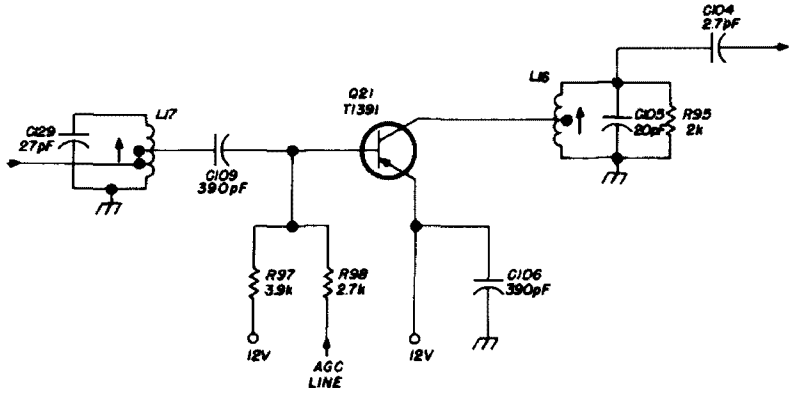
fig. 7. Special i-f circuit passes signal in two directions.



through resistor R23, and is grounded directly for rf by C19. The emitter is fed dc from the 12-volt line through load resistor R24. C131 eliminates any rf that is "hanging on," and C20 couples the audio to the volume control. Base bias is developed by divider R20-R21.

through R124 to the base (through R121) and the emitter (through R123). The rf gain control (R122) forms the bottom part of the divider from the 100-volt negative line. All this makes the collector positive with respect to emitter.

fig. 8. Rf stage uses tapped coils to match transistor impedances.



agc amplifier

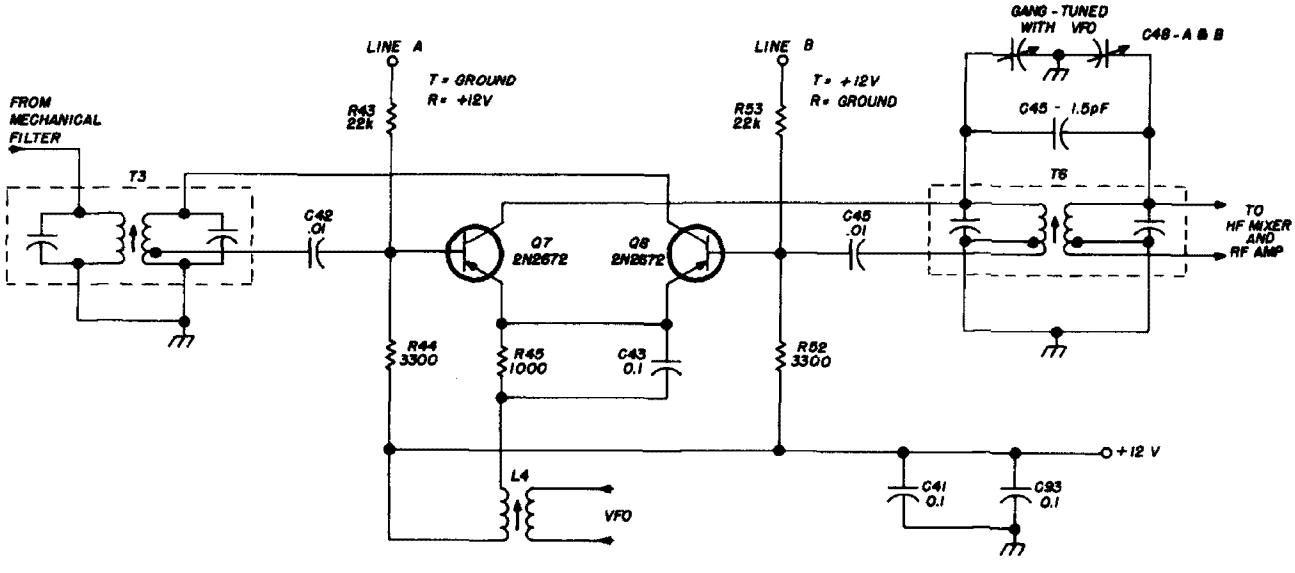
One transistor agc circuit, from the Galaxy V Mark 2, is shown in **fig. 11.** Audio signals are fed through C137 from the product detector to the base of Q8. This is an npn stage which requires positive collector voltage. But the collector must develop a negative voltage to apply to vacuum-tube grids (the set is a hybrid).

If the collector must be somewhat negative with respect to ground, the emitter and base must be even more negative. As you can see, a 100-volt negative line is tied

Q8 rectifies the audio signal and develops a negative dc output across the R125, R126, and part of R122. C138 and C139 bypass what audio is left, and the negative dc voltage is used to bias the grids of the rf and i-f amplifiers.

Manual gain control for both rf and i-f stages is provided by R122. The arm taps off a portion of the voltage from the negative 110-volt line, and applies it to the grids of the rf and i-f amplifiers through R126. Only if the dc output of Q8 exceeds the amount set by R122 does the agc take effect.

fig. 9. Two-directional mixer stage with common oscillator for both receive and transmit.



bias regulator

National's NCL 2000 is a linear power amplifier in which a pair of power tubes in parallel furnish 2000 watts PEP when driven by an ssb exciter. For proper sideband operation, the PA grid bias must be held constant despite varying line voltage or load changes. **Fig. 12** shows the circuit that accomplishes this task.

A dc supply furnishes 90 volts negative directly to the collector of the regulator transistor. This —90 volts is also tied, through R39, to the emitter of Q2, the error amplifier. Q2's emitter is held at —18 volts by zener diode D9, as a steady reference. C17 and C12 bypass any transient voltages that might be too quick for the zener diode.

The collector of Q2 is also tied to the 90-volt negative line, through R38. C6 and C8 bypass any transients that get into this branch. Q2's collector is more negative than its emitter or its base (which goes to ground through R42). Q2 is biased on and conducts.

But Q2's collector is tied directly to the base of regulator transistor Q1. Also, remember, Q1's collector is tied directly to the 90-volt negative line. Q1's emitter, which is the output element of the regulator, is returned to ground through resistors R40, R41, and R42—a voltage-sensing divider across the regulator output line. C11, C15, C16, C10, and C49 bypass any transient voltages in the grid-bias sensing network.

When Q2 is on, Q1 is also on, but its base bias depends on the conduction of Q2. Current flows from the 90-volt negative line through Q1, and through the sensing divider R40-R41-R42 to ground. This places Q1's emitter—and the PA grid-bias line—normally somewhere between —25 and —45 volts with respect to ground.

If the voltage on the grid-bias line changes, it affects voltage at Q2's base. Q2 amplifies this change and inverts it at the collector.

fig. 10. Emitter-follower product detector.

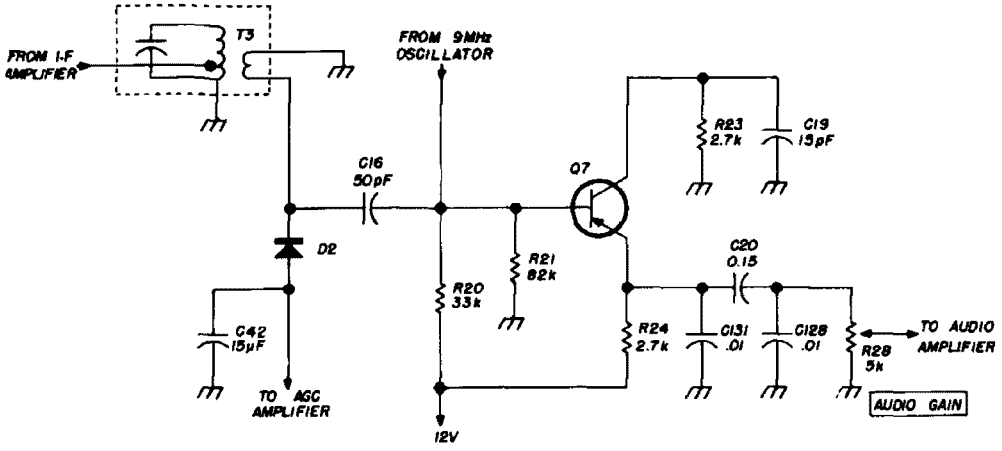
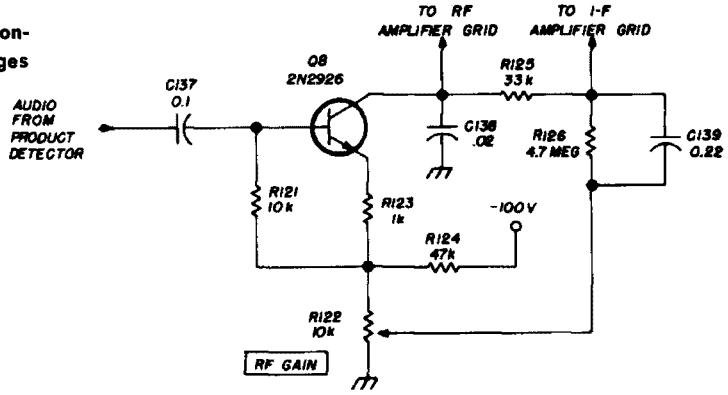
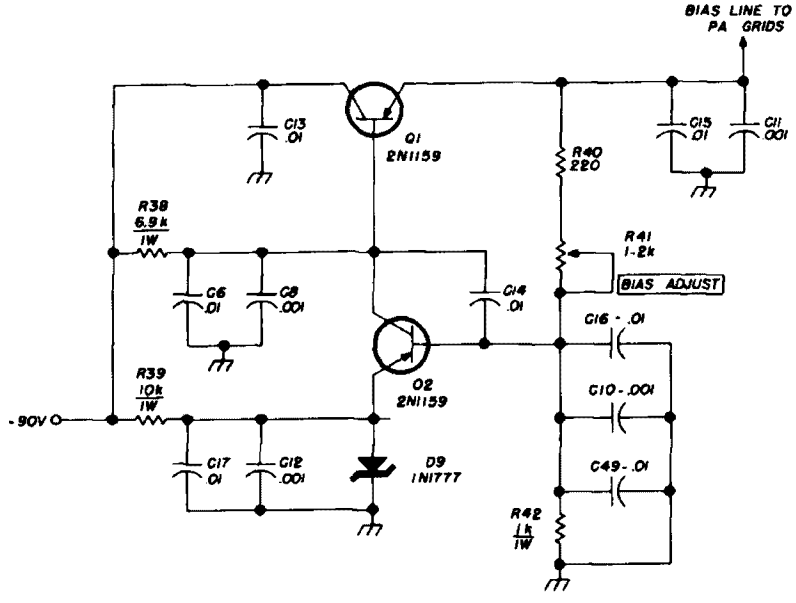


fig. 11. Solid-state automatic gain control circuit that furnishes control voltages for vacuum tubes.



Thus, any error signal is applied to the base of Q1. The emitter-collector conductance of Q1 changes and restores the original voltage at Q1's emitter. The normal bias value is adjusted manually by R41, which sets the operating point for the base of Q2.

fig. 12. Two-transistor circuit regulates power-tube grid bias.



future of transistors in ssb

That's a real handful of transistor circuits. Most sideband functions, you have seen, can be performed by today's transistors or diodes. You may have noticed that no rf power stages were mentioned here. This lack will soon be filled. New silicon "overlay" transistors promise up to several hundred watts and can work beyond 50 MHz. Some are available now, but at high cost. In time, though, all but the very highest-powered ssb transmitters will be completely solid-state.

ham radio

drive indicator

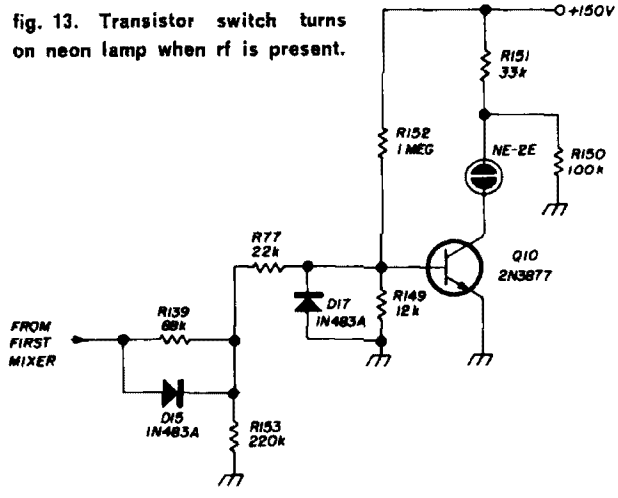
The Drake R-4B receiver is often used with a companion transmitter. When it is, the receiver vfo is used to excite the transmitter. The frequency is then identical for receiving and transmitting.

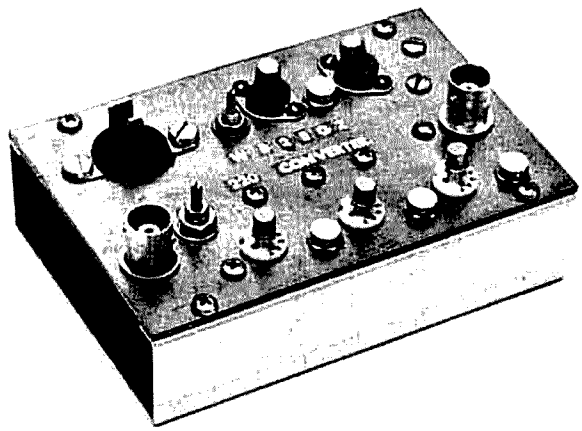
But if the transmitter were fired up without rf drive, the final tubes could be damaged. A neon lamp might be used to monitor rf drive directly, but the rf level is too low to fire a neon. So, a transistor switcher is used instead, as shown in **fig. 13.**

R150 and R151 form a divider from the 150-volt line to ground. Q10 is an npn grounded-emitter transistor, and is not normally conducting, so it has a high emitter-collector resistance. The lamp stays off.

When the vfo operates, rf from the first mixer stage is rectified by D15; a positive voltage develops across R153. This dc is applied through R77 to the base of Q10. (D17 is a protective diode which shunts any negative peaks to ground.) The transistor is biased into conduction and becomes a low resistance from its collector to ground. The neon fires.

fig. 13. Transistor switch turns on neon lamp when rf is present.





mosfet converter for 220 mhz

You must still select these semiconductors for the higher frequencies—but it's worth it

Don Nelson, WB2EGZ, 9 Green Ridge Road, Asland, New Jersey 08034

Superior gain characteristics of the metal-oxide semiconductor field-effect transistor (MOSFET) have been exploited in ham-band converters for six and two meters in **ham radio's** June and August 1968 issues.[1,2] Although its gain diminishes slightly between 200 MHz and 300 MHz, the MOSFET—particularly the dual-gate type—is still a fine performer at 220 MHz. The deluxe combination of low noise, high gain and low cross-modulation are packaged in the following solid-state converter, which is reasonably easy to construct and tune.

circuit considerations

The 1¼-meter band is more demanding of semiconductors than are the lower frequencies. Only by selecting high-quality components was it possible to achieve the desired level of performance. Much to my surprise, the oscillator and multiplier transistors are equally as critical as those used in rf stages.

The first rf amplifier uses an RCA 3N159 dual gate MOSFET, which is a low-noise version of the RCA 3N140. In using the 3N159, you achieve the lowest possible noise figure, about 2.5 dB. Very low feedback capacitance

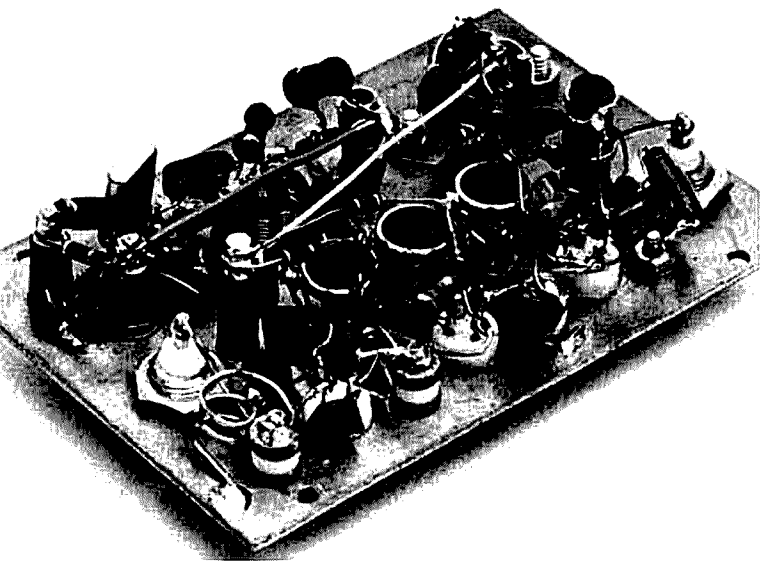
—less than 0.02 pF—allows operation with no neutralization. The second stage, being less critical, uses the 3N140 operating at lower gain to prevent overload.

Two rf stages may be stagger tuned if wideband operation is desired. Unlike the two-meter converter, which had to be stagger tuned, this converter can have both stages peaked on the same frequency. This added utility is made possible by using dual-gate MOSFET's in both rf stages, with Teflon sockets, to minimize feedback capacitance.

Only one change is made in the 3N141 mixer circuit from that in the 6- and 2-meter converters. It is the addition of a neutralizing circuit from gate 2 to ground.[3] The L and C resonate at 14 MHz, effectively bypassing gate 2 at the intermediate frequency. (Values must be changed if output other than 14 MHz is employed.) Incoming signals will be improved one S-unit or more with this circuit. The technique may also be used on the two-meter converters.

Birdies and "garbage" on 220 have not helped to make it a popular band. One way to improve the situation is to use a high-frequency local oscillator as was done here. A 2N3478 drives a 103-MHz fifth overtone crystal whose frequency is doubled in a second 2N3478. Cheaper transistors did not deliver enough injection voltage for good gain in the mixer.

I find that the use of back-to-back diodes in the antenna circuit is quite controversial.



Some feel they are essential to prevent burnout of the front-end transistors, while others say this is not so.* My friend Ed, W2DMR, uses a 3N140 preamplifier with no diodes on ten meters with a kilowatt. His only isolation is a DK60G relay. He operates within 1/4 mile of another kilowatt on the same band without problems. I have removed the diodes in all my converters without mishap, but 10 watts of rf is the greatest power I've generated on any band. Diodes are shown in the schematic for those who care to use them.

Sometimes rf burnout has been traced to slow decay of the transmitter's output, rather than the lack of isolation of the relay. In this case, large amounts of rf may load the converter during coaxial relay switching. A very good solution to this problem is found in delaying the relay, or by using two sequentially operated relays—one for the transmitter; one for the receiver.

construction

As with the 6- and 2-meter converters, all components were mounted on copper-clad printed-circuit board according to the layout shown. The board is fastened to a Bud CB1626 chassis. Satisfactory operation is once again possible without shields. For those who may have forgotten the rules for handling MOSFET's, the following should be observed:

1. Keep MOSFET leads shorted until ready to use. (These devices are shipped this way.)

2. When cutting leads, grasp lead and case with fingers to reduce possibility of electrical and mechanical shock.

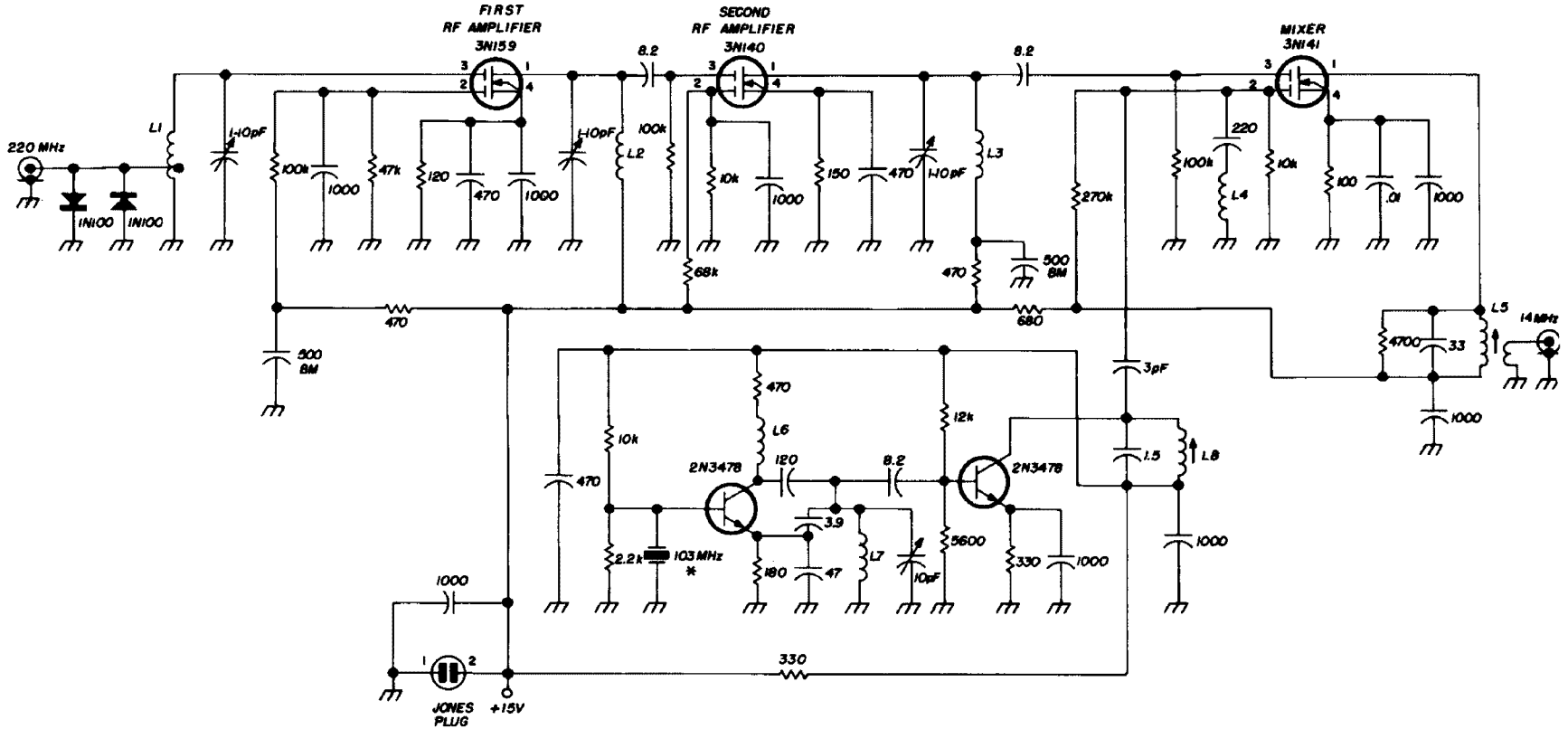
3. Do not solder or change components with MOSFET's in their sockets. (Such a practice may be acceptable if you use a soldering iron with a grounding system.)

4. Never insert or remove transistors when power has been applied.

performance plus

A comparison of this converter was made with my Nuvistor converter whose noise

* Depends on the transistors. Quality control and inspection mean different things to different manufacturers. Better use the diodes and be safe. **Editor.**



L1 2½ turns number 20, spacewound, ¼" diameter; tapped 1 turn from ground

L2, L3 2½ turns number 22, ¼" diameter

L4 0.68 μH. (J. W. Miller 9310-08)

L5 3 μH. 17 turns number 22 on 3/8" diameter slug-tuned form (J. W. Miller 20A336); secondary is 3 turns number 26 on cold end of L5

L6 1.5 μH (J. W. Miller 9310-16)

L7 0.22 μH (J. W. Miller 9310-02)

L8 3½ turns number 22 on .205" diameter slug-tuned form, red core (J. W. Miller 4300-3)

103-MHz crystal is 5th overtone (International Crystal FM-1-GP)

All 10-pF trimmers are JFD type VAM-010 (Allied Radio 43F345)*

fig. 1. Schematic diagram of the 220-MHz converter. Capacitors marked BM are button micas; bypass capacitors are discs or laminated ceramics.

*Available from Allied Radio Corporation, 100 N. Western Avenue, Chicago, Illinois 60680. Order catalog number 43F3475. $3.95 each plus postage: shipping weight, 2 ounces.

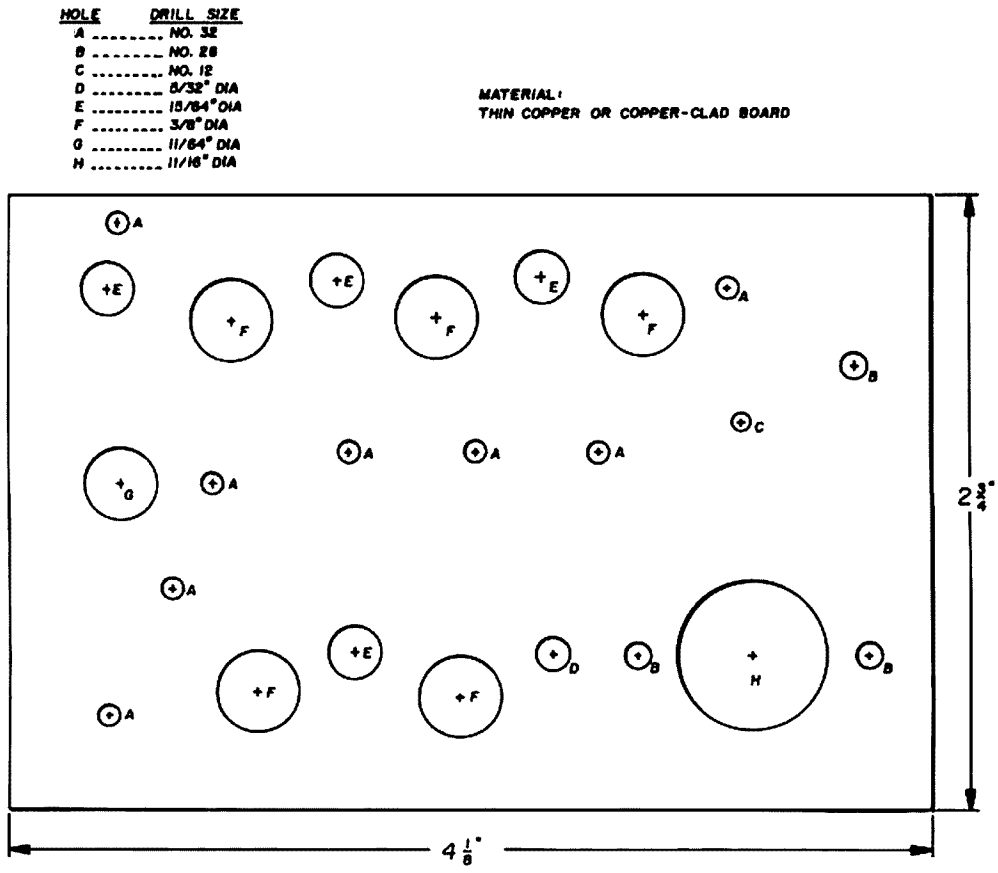


fig. 2. Chassis layout for the 220-MHz mosfet converter; this chassis designed for Bud CB-1626 chassis.*

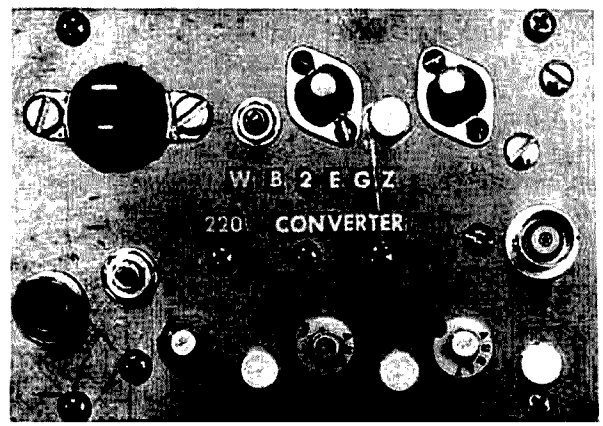
figure is about 5 dB. The noticeably quieter MOSFET's are music to my ears. I find that the gain is comparable, which is not an easy achievement with the best of the current JFET's.

While 220 is an orphan band in many parts of the U.S., the Mt. Airy VHF Society has kept it alive in the Philadelphia area. Only one station has managed to overload my MOSFET converter. (He is hard on the tubes too!) Admittedly, this is no test for comparison, but I feel the MOSFET is at least as good as the proven two-rf-stage 2-meter converter[2] with respect to cross modulation and dynamic range.

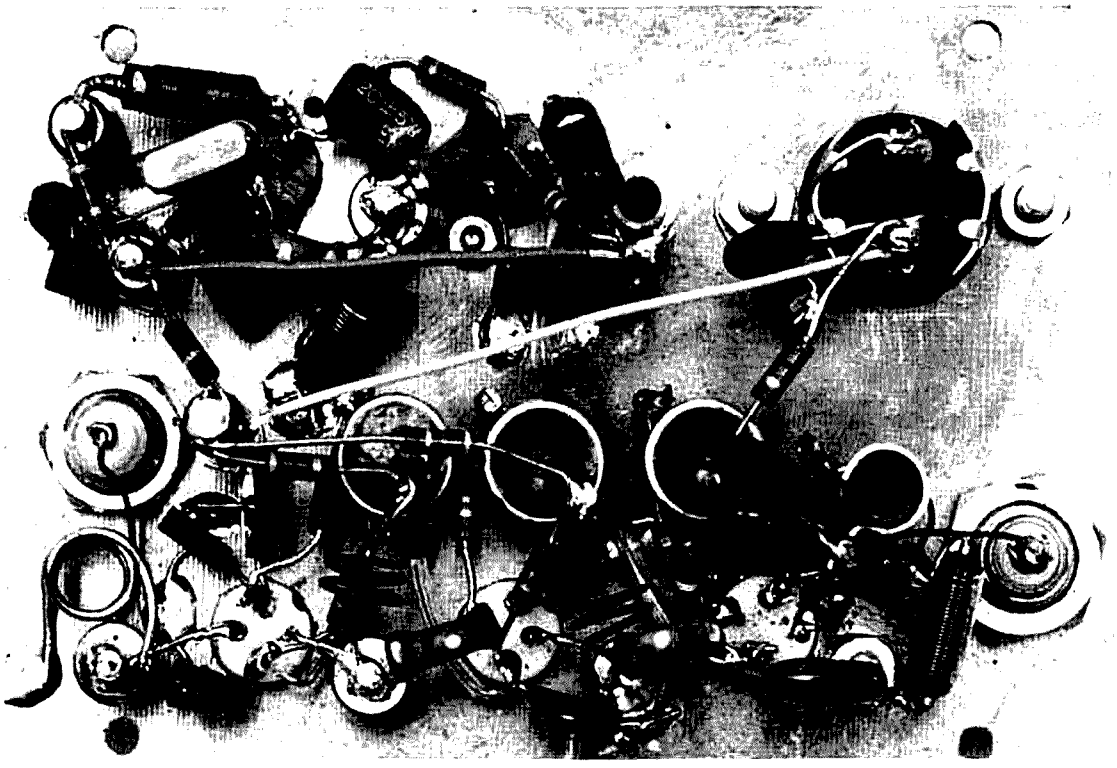
mosfet's and the future

Only a few of the semiconductor manufacturers are manufacturing MOSFET's, and not all are interested in the depletion-mode types for vhf such as the RCA transistors used in this series of three articles. Stability problems of the oxide layer which once plagued the MOSFET have been largely solved, giving time for curing the ills of static burnout and frequency limitations.

MOSFET amplifiers at 500 MHz are practical now, but high-volume production is not in effect on these units. Some feel the



*Bud CB-1626 chassis available from Allied Radio Corporation, 100 N. Western Avenue, Chicago, Illinois 60680. Order catalog number 42E7812, $.55 plus postage; shipping weight, 12 ounces.



Under-chassis construction of the 220-MHz mosfet converter.

upper frequency limit of the MOSFET is 1000 MHz, but science may soon disprove that notion. Even now, Motorola has incorporated a layer of silicon nitride to eliminate static burnout of the oxide layer.

Fairchild has integrated additional circuitry in the MOSFET package to reduce the burnout hazard, while other manufacturers are experimenting with still different techniques. Presently, there is no simple solution compatible with all MOSFET's, but the fact also may be changed in the near future.

The metal-oxide semiconductor field-effect transistor answers the challenges of high input impedance, extremely low feedback capacitance, high gain, low noise, inexpensive construction, wide dynamic range and low cross modulation. It is so distinctive that it invites comparison in all rf and many dc applications.

references

1. D. W. Nelson, "The WB2EGZ Six-Meter MOSFET Converter," *ham radio*, June, 1968, p. 22.
2. D. W. Nelson, "The Two-Meter Winner," *ham radio*, August, 1968, p. 22.
3. H. M. Kleinman, "Application of Dual-Gate MOS Field Effect Transistors in Practical Radio Receivers," *IEEE Transactions on Broadcast and TV Receivers*, July, 1967.

ham radio

next month in ham radio magazine:

signal detection in the presence of noise

programmable repeater identifier

stable transistor vfo

universal solid-state preamp

power supplies for ssb

analyzing incorrect dc voltages

sloping dipole antenna

plus many more...

stub-switched stub-matched antennas

A simple multiband antenna system that features automatic stub matching as you change bands

John J. Schultz, W2EEY, 40 Rossie Street, Mystic, Connecticut 06355

Transmission lines can be used for various purposes besides transferring power to an antenna. Depending on the electrical length of the line, it can function as a frequency-dependent switch, an impedance-matching transformer or a reactive circuit element. You can devise a number of interesting antenna designs with the first two functions. Normally these characteristics are used separately, but there is no reason why they can't be combined.

Before you digest any designs, it's important to have a clear idea of how transmission-line sections work as impedance-matching transformers and frequency-dependent switches.

transmission line characteristics

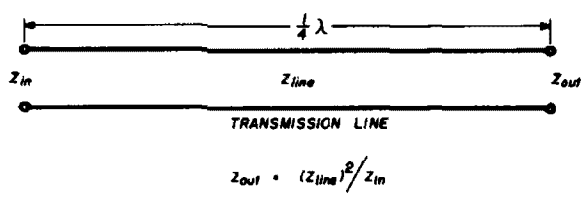
A lossless transmission line one-quarter wavelength long will transform the impedance at its input terminals to an impedance at its output terminals equal to

$$\frac{(Zo)^2}{Zin}$$

where Zo is equal to the impedance of the line.

However, this is only true when the input

fig. 1. Basic quarter-wave transmission-line transformer; limiting cases occur where input impedance is either zero or infinite.

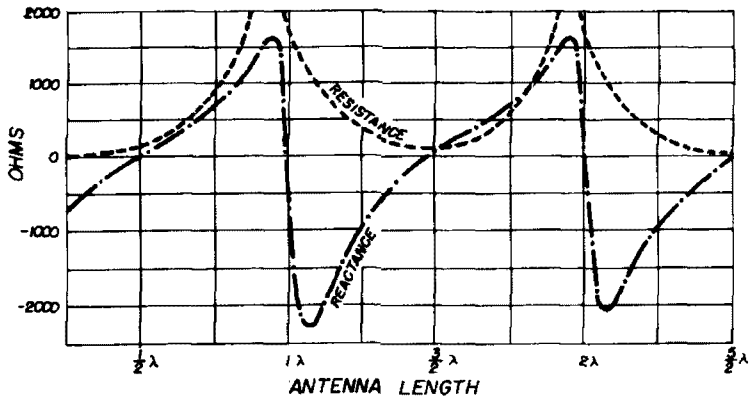


or output impedance is a pure resistance. Although the whys of this impedance transformation can be shown mathematically, the proof is somewhat tedious. Perhaps the simplest way to visualize the action is to remember that the input voltage and current vectors undergo a 90° phase shift on the quarter-wave line. Therefore, their relative amplitude values are reversed. If the input impedance is lower than the line impedance, the output impedance is always higher. If the input impedance is greater than the line impedance, the output impedance is always lower.

Any number of quarter-wave sections can theoretically be used in series if one particular line does not provide the desired transformation. If the transmission line is one-half wavelength long, the output impedance is the same as the input impedance since this is the same as putting two quarter-wave sections back-to-back.

The limiting case occurs as shown in **fig. 1** when the input impedance is either zero (short circuit) or infinite (open circuit). It can be seen from the impedance-transformation formula that the output impedance must be opposite from the input impedance. With this in mind, we can use the quarter-wave transmission line as an rf switch.

fig. 2. A dipole antenna has reactive and resistive components at harmonic frequencies as shown here.



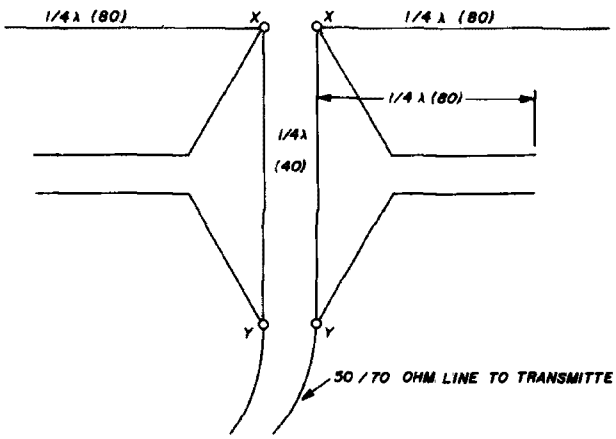
However, the switch is frequency dependent. If the input impedance is zero at a frequency where the transmission line is a quarter-wavelength long, the output impedance is infinite and looks like an open circuit. If line length is fixed and the frequency is doubled, the transmission line represents one-half wavelength and the output impedance is the same as the input impedance.

Therefore, by controlling the termination and length of the transmission line, it can be used as an impedance transformer, a 1:1 impedance transfer element or a frequency-dependent switch.

dipole harmonic operation

A simple resonant half-wave dipole presents a pure resistive impedance which matches 50- or 70-ohm cable fairly well. If the dipole is used at harmonic frequencies, its terminal resistance and reactance will vary as shown in **fig. 2.** At even multiples of a half-wavelength, the terminal impedance is highly resistive with practically no reactive component; at odd multiples, the terminal impedance is resistive at almost the same value as at the fundamental frequency with no appreciable reactive component. This is the reason you can use a 7-MHz dipole successfully on 15 meters with the same feedline.

fig. 3. A stub-switched and -matched antenna for 80 and 40; construction is shown in fig. 6.



To feed maximum power into a dipole at any multiple frequency of its fundamental, you have to match the antenna's terminal impedance. The usual way to do this is to use a resonant transmission line and an antenna coupler at the transmitter. By using combinations of quarter-wavelength transmission line sections at the antenna, however, it's

possible to obtain multiband operation without an antenna coupler or any tuning or bandswitching circuits. The transmission line going to the transmitter will still operate at a low swr on each band.

some basic antennas

Once you understand how the transmission line sections can be used, you'll undoubtedly be able to come up with designs of your own. As a starter, here are some basic designs I have investigated.

A dipole which automatically bandswitches from 80 to 40 meters is shown in **fig. 3.** On 80 meters, the open-circuited quarter-wavelength sections (shown horizontally) between points X and Y effectively short out the vertical quarter-wavelength section. This effectively connects points X and Y together as in a normally fed dipole.

On 40 meters, the flat-top portion of the antenna is a full wavelength long with a rise in input impedance as shown in **fig. 2.** The horizontal transmission line section which was one-quarter wavelength long on 80 meters is one-half wavelength long on 40 and presents an open circuit between terminals X and Y.

The vertical quarter-wavelength section is effectively connected between terminals X and Y; this section acts as a transformer, and its characteristic impedance is chosen to match the transmission line to the transmitter.

Normally, if 50- or 70-ohm coaxial cable is used, the matching section can be made from 300-ohm twinlead. The characteristic impedance of the horizontal transmission line sections aren't particularly important since they only perform a switching function.

The antenna operates as a normal dipole on 40 meters; the main radiation remains broadside to the line of the flattop with about 1.8-dB gain.

Variations of the same antenna for different bands are shown in **fig. 4.** The antenna of **fig. 4A** is designed for 20 and 10 meters

fig. 5. Using transmission-line sections for switching vertical antennas.

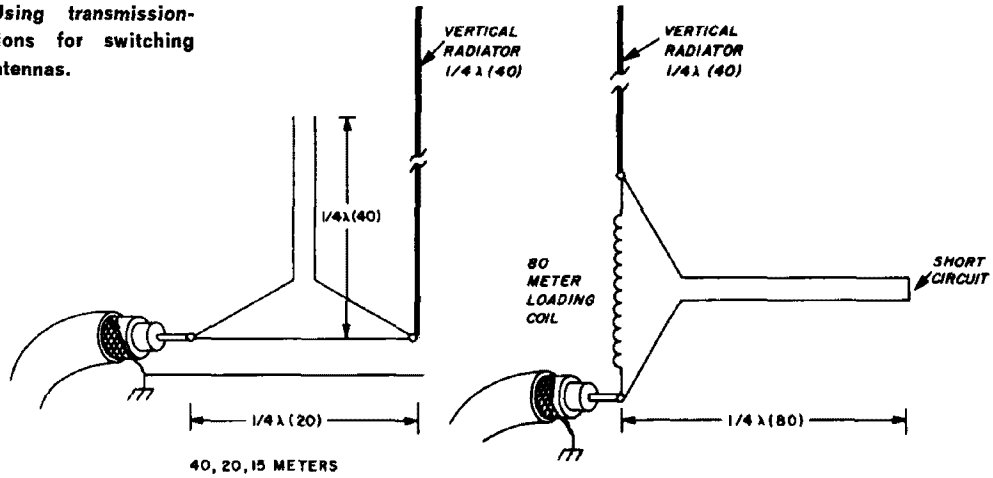
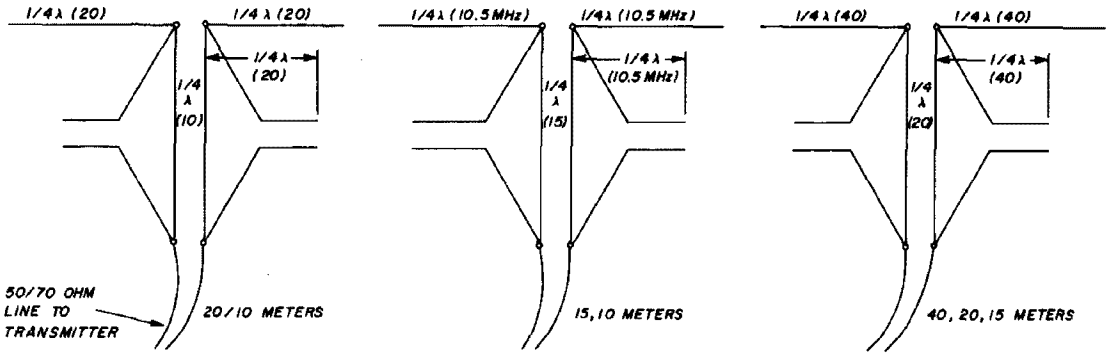


fig. 4. Variations of the antenna of fig. 3 for other bands.



and is simply a scaled version of the antenna shown in **fig. 3.** The antenna of **fig. 4B** is for 15 and 10 meters but is a bit different. Since you can't cut a half-wave dipole for 15 meters and use it as a full-wave dipole on 10, a different approach is used. The dipole is cut one-half wavelength long on 10.5 MHz; this makes it a full-wave dipole on 15 and approximately three half waves long on 10.

If you analyze the switching section, you'll find it produces a short circuit on 10.5 MHz, an open circuit on 15 meters and a short circuit again on 10 meters. The antenna is matched on 10.5 MHz and while this feature has no amateur value, it may be useful for WWV reception. Because of the length of the flattop, the antenna has about 1.8-dB on 15 and 10 meters. However, on 10 meters the radiation pattern takes the form of a cloverleaf characteristic of a 3/2-wave dipole.

The antenna shown in **fig. 4C** is a little more conventional although it is basically the same as **fig. 3.** Operation is possible on three bands —40, 20 and 15 meters—because the flattop is 3/2-wave long on 15 and the transmission-line switch shorts out the matching section.

The use of transmission line sections doesn't have to be limited to dipoles; they can be used just as easily with unbalanced vertical antennas. The antenna in **fig. 5A** is basically one half the antenna shown in **fig. 4C.** Although it resonates on 40, 20 and 15 meters, it is mainly useful on 40 and 20 since the radiation angle is quite high on 15 meters. This is because the vertical radiator exceeds 0.64 wavelength. A simple two-band antenna can also be built as shown in **fig. 5B;** in this design the transmission line section acts as a switch.

construction

You should observe a few precautions when building any antenna. For one thing, the physical length of the transmission line must take the velocity factor of the line into consideration (typically 70% for 300 ohm twinlead). The line sections shouldn't be formed into coils and placed at the center of the flattop since this will introduce spurious resonances. It's better to bring both ends of the sections to the center of the antenna in the form of a drooping "U" as shown in **fig. 6** and then make all the interconnections.

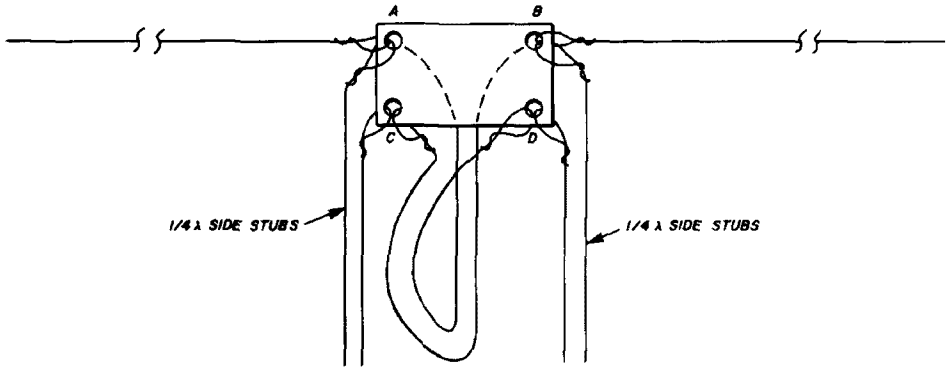


fig. 6. Construction of the stub-switched antenna. The center stub section is looped between terminals AB and CE. The coaxial feedline is connected to terminals C and D.

The values shown in **fig. 2** are typical for wire antennas made from number 12 or 14 wire. If the antenna is made from very thin wire and used on 80 meters, the impedance values at harmonic frequencies may be higher than those shown. In this case it's a good idea to put the antenna together temporarily without the switching sections and operate it on the harmonic frequency; then measure the swr to determine if you've picked the right impedance for the matching section.

You should have no difficulty with the switching sections if they are cut to formula length. The power handling capability of the antenna is determined by the rating of the transmission line used for the line sections.

ham radio

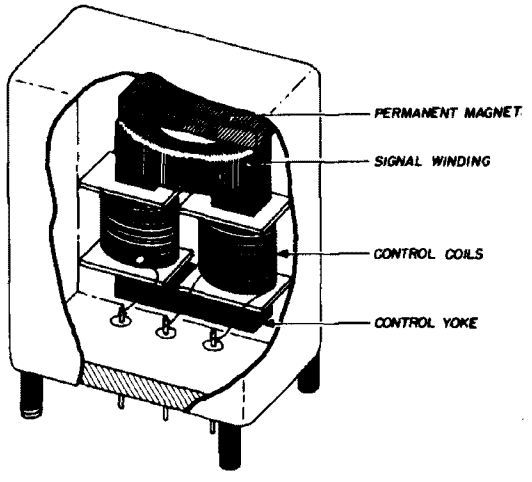
solid-state current-controlled tuning

How to use a new electronic component that adds versatility to a number of circuits

Robert M. Brown, K2ZSQ, 5611 Middaugh Avenue, Downers Grove, Illinois 60515

Coming for the most part as fallout from military research and development—where they have been used for almost ten years—are a fascinating line of variable inductors. These inductors are electrically variable, but unlike varicaps, they alter the inductance of a circuit as a control current is varied from 0 to 60 mA.

Cutaway of a typical Vari-L current-controllable inductor.

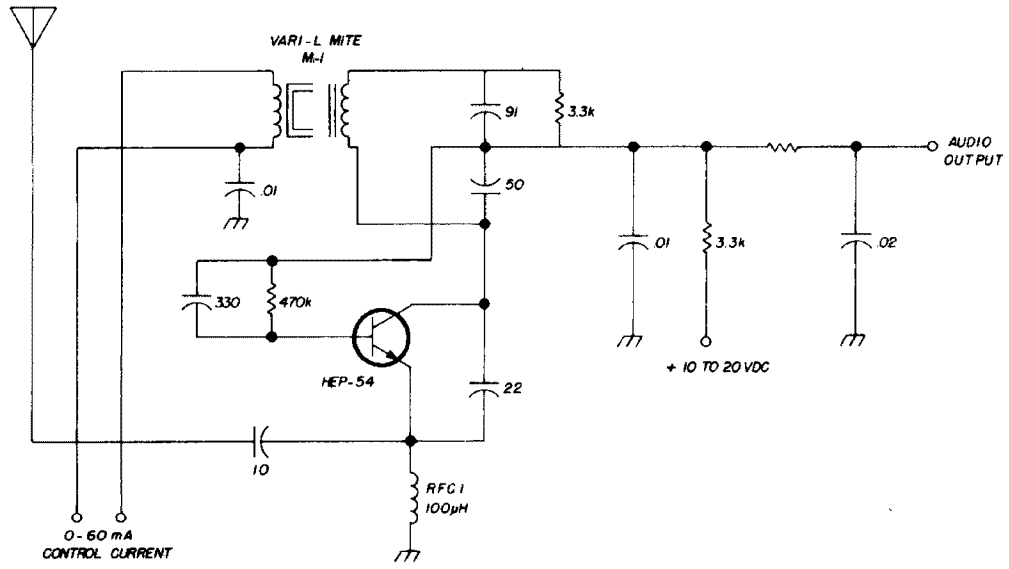


One of the circuits developed by the Vari-L Company* to demonstrate typical applications of the Mite series variable inductors is shown in **fig. 1.** As the control current is varied from 0-60 mA, this superregenerative receiver's detector tunes from 20 to 50 MHz. The audio output can be fed to a simple transistor amplifier. While the inductor shown presently sells for $15.80, the manufacturer expects production costs to fall appreciably as time goes on.

(at zero control current). With the 100-mH units, frequency bands of 3 through 30 kHz can be covered; with the .05 μh inductor, a 250-300 MHz range can be realized.

When you use these inductors, it's recommended that the signal winding be connected to the tuned circuit the same way as any conventional high-Q coil. Next, it is resonated with a suitable tuning capacitor, and the control winding is connected to a source of continuously-variable current (such as a battery and a rheostat).

fig. 1. Superregenerative receiver using a Vari-L coil. Receiver tunes from 20 to 50 MHz as the control current is varied from zero to 60 mA.



how the inductor works

These tiny variable inductors are two-part assemblies. The control winding is wound on the support legs of a U-shaped ferrite or laminated core. The signal winding is wound on a ferrite core positioned over the open end of the U; this winding is used as a tunable coil in a Colpitts oscillator or other two-terminal networks. When ac or dc current is passed through the control winding, it causes the magnetic flux to vary—decreasing the over-all permeability of the signal-winding core.

These Mite-series units are available with inductances from 100-mH to .05 μH

* Vari-L Company, Inc., 207 Greenwich Avenue, Stamford, Connecticut 06904.

By varying the current, the operating frequency is smoothly controlled. A tracked chain of several units can be controlled equally well—tuned to the same or harmonically related frequencies. The characteristic curve of a typical Mite inductor is shown in **fig. 2.**

applications

Since the inductance can be varied slowly or at high speed rates—up to several million times per second—command can come from just about anywhere. It can be a manually-operated source, an electronic circuit or a transducer. They are useful for remote control or closed-loop systems, for sweeping oscillators or passive filters and

for frequency switching.

These variable inductors have been used in receivers, both in local oscillators and gang-tuned rf amplifiers, sweep generators of all types from audio to vhf, frequency modulators for transmitters, ultrasonic test generators and spectrum analyzers. Since the unit is sealed, ruggedly built and without any moving parts, field failures are unknown. Early versions are still going strong after 12 years according to the manufacturer.

Additionally, Vari-L is now concentrating on the development of new uhf variable-inductor designs and resonant-cavity units which will extend the useful region of the devices into the gigahertz range. This should be of particular interest to uhf enthusiasts looking for better ways to remotely control their plumbing installations, such as preamps/converters that are employed at the feedpoints of parabolic dishes and corner reflectors.

One vhf application is shown in **fig. 3.**

fig. 3. Oscillator that tunes from 170 to 240 MHz.

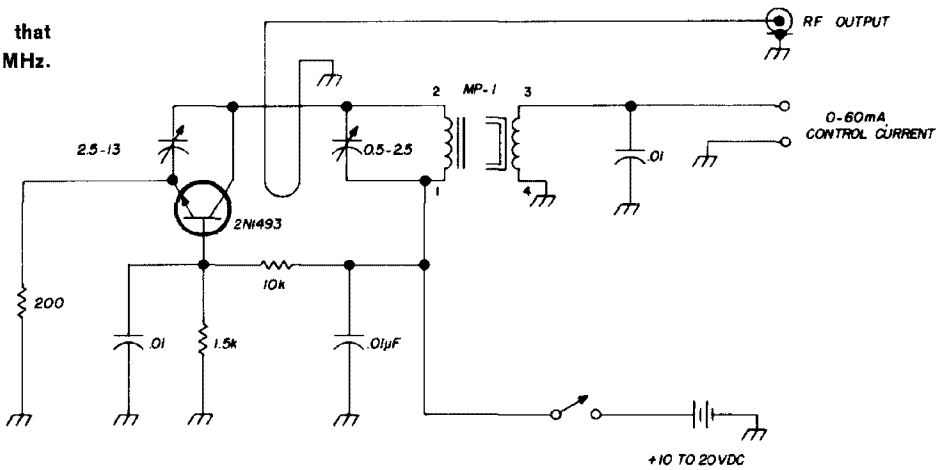
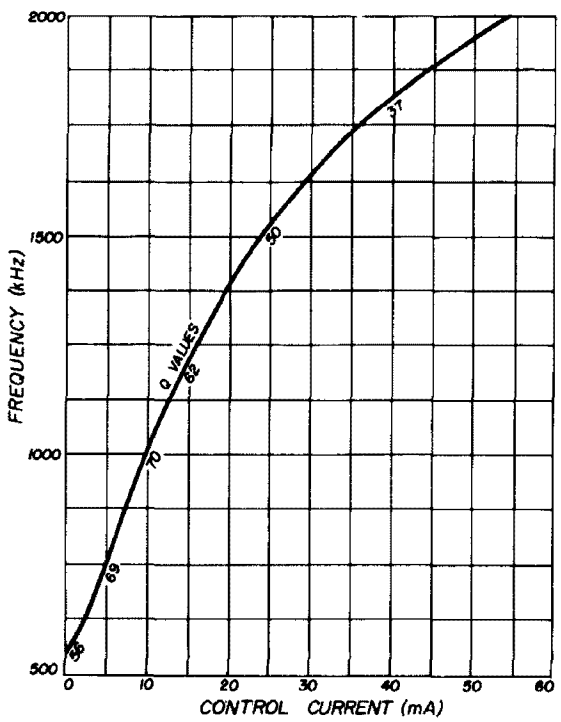


fig. 2. Characteristics of a typical Vari-L inductor.



This oscillator tunes from 170 to 240 MHz. The frequency is varied by the 60-mA control current fed to the MP-1 inductor at terminal three. The output link is a conventional hairpin located close to the tank and adjusted for maximum output at the frequency desired. The 2.5-pF variable capacitor may be a fixed value if only 220-MHz operation is desired. Do not substitute other transistors for the 2N1493.

Another interesting application is shown in **fig. 4.** This circuit provides one milliwatt into a 50-ohm load in the 260- to 380-MHz range. A Fairchild 2N3563 is recommended by the manufacturer, although a 2N918 can be substituted with good results using the same circuit constants. Usual construction techniques are recommended to insure oscillation over the entire frequency range. Further, a shielding can is recommended for the Mite MP-1.

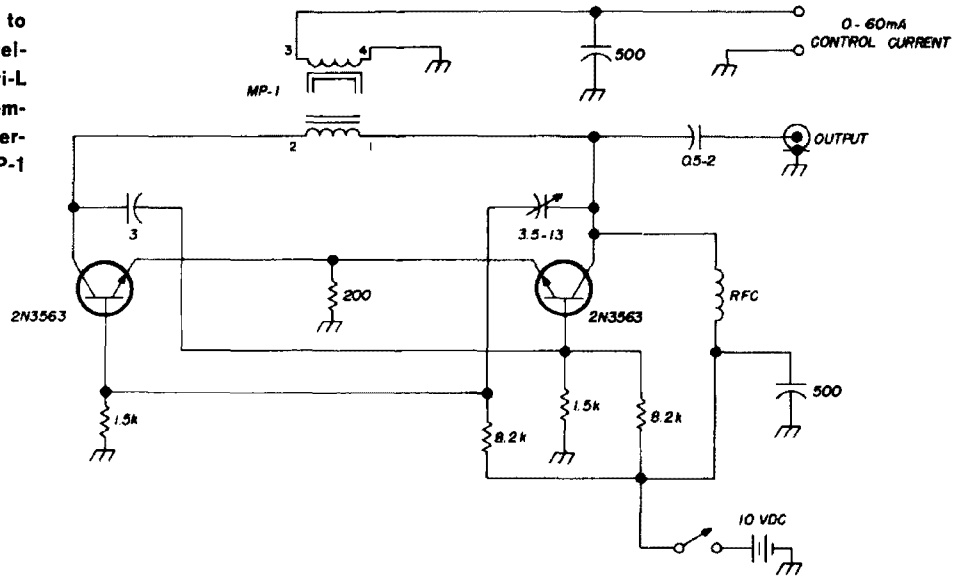


fig. 4. A 260- to 380-MHz vfo developed by Vari-L engineers to demonstrate the versatility of the MP-1 inductor.

Lastly, **fig. 5** shows how a Vari-L MK-100 inductor can be used in a 5 to 75-kHz continuously variable oscillator. With the 2N2712 transistors shown, output will be on the order of 0.1-volt rms with three-percent distortion. Unlike previous examples, this circuit requires a somewhat lower control-current range—zero to 15 mA. Other transistor types can be used with good results, including the 2N4124, 2N4265, HEP-54 and GE-10.

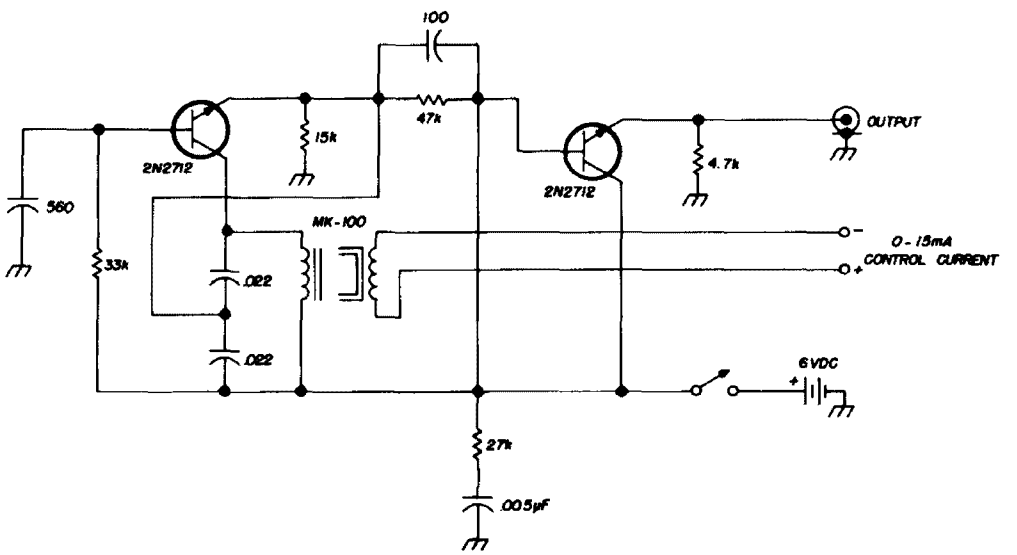
In conclusion, it should be noted that Vari-L products are presently not widely available at electronic distributors. If you're interested, it's suggested that you contact the manufacturer directly for literature. Vari-L will respond to interested amateurs and will ship single-unit quantities by first-class mail.

references

1. "Remote-Tuned Superregenerator," *Radio-Electronics*, May, 1967, p. 53.
2. "Manufacturers Circuit," *Popular Electronics*, June, 1967, p. 77. Description of a "Z-Match" unit also manufactured by Vari-L, showing application in a broad-band amplifier circuit. Readers interested in the wide-band transformer should request Bulletin 567.
3. "Vari-L Condensed Catalogue: Electrically-Variable Inductors, Sweep Oscillators, Develoboards," available on request from Vari-L Co., Inc.

ham radio

fig. 5. Variable oscillator that tunes from 5 to 75 kHz.



some notes on cubical quad measurements

L. W. VanSlyck, W4YM, 406 Bon Air Avenue, Temple Terrace, Florida 33617

Much has been written, and probably much remains to be written, about cubical quad antennas. A review of measurement methods for these popular antennas seems in order for those desiring to improve the performance of their quads and for those who wish to build one from scratch.

The basic beam consists of a driven element and reflector. Measurement data are given here for arrays with n-elements, however, for those interested in a larger antenna.

Test equipment includes:

1. Accurately calibrated receiver
2. Grid-dip oscillator
3. Antennascope (rf bridge)
4. Swr bridge
5. Signal source capable of providing a few watts at the frequencies of interest.

The accurately calibrated receiver is used to check the grid-dip oscillator to obtain a frequency reading within one-half percent. The signal source may be your transmitter or transceiver if power can be reduced to a few watts.

driven element resonant frequency

First, disable all elements except the driven element. Open these elements and fold the wire back over itself. If you don't do this, you'll get several readings on the GDO when it's coupled to the driven element. Now loosely couple the GDO by using a small "gimmick" made from a male coax connector and a short loop of wire. Solder one end of the wire to the connector shell, and connect the other end to the center pin, making a loop about a half-inch in diameter.

The gimmick replaces the coax on the driven element for the measurement. Next, raise the antenna as high as you can conveniently reach to make the measurement. The plane of the antenna should be perpendicular to the ground. Find the resonant point, using the GDO with very loose coupling, then check the GDO with your calibrated receiver. This should give an accurate measurement of the driven element resonant frequency. If there are other elements to be measured, proceed in the same manner. Keep a record of the data.

Cubical quad driven elements are cut according to the relationship

$$L = 1000/f$$

where

L is the length of wire (feet)
f is the resonant frequency (MHz)

Some manufacturers use a constant of 1005, but the difference, in terms of total percent length, is insignificant for amateur work.

On 20 meters, the wire length should be changed 5.7 inches for each frequency change

of 100 kHz. In other words, if your driven element resonates at 14.2 MHz, and you want it to resonate at 14.3 MHz, subtract 5.7 inches from the total length, or about 1.5 inches from each side ($4 \times 1.5 = 6$, which is close enough).

reflector resonant frequency

If coils are used in the reflector to lower its frequency, simply couple the GDO loosely to the coil, and measure f; it should be 3 to 5 percent lower than the driven element at the point chosen for it's operating frequency. If a stub is used to lower the frequency, couple the GDO to the stub. If you can't get a satisfactory reading by this method, then coil up the wire in the stub to make a small loop, and couple the GDO to the loop. The loop should be a single turn, or even a half turn, to get the correct resonant frequency. If f is obtained by increasing the wire length 3 to 5 percent, follow the procedure given above.

director resonant frequency

An identical procedure is used for checking director resonant frequency. These elements should resonate from 3 to 5 percent higher in frequency than that of the driven element. (I use 5 percent.) It's possible, if more than one director is used, to resonate the first director at 5 percent, the second at 4 percent, etc. No data have been obtained concerning shortening each director from the previous one by a fixed amount, but yagis are often constructed in this manner, so the same reasoning might apply to a quad.

input impedance

Couple the antennascope or rf bridge to the GDO, and couple the output of the rf bridge to the driven element. Set the GDO at the desired frequency, and tune for a minimum dip on the rf bridge. Next, read the input impedance on the rf bridge. A curve may be plotted, if desired, of frequency versus input impedance. It will give you a good idea on how flat the line really is. Remember, loose coupling is used in all measurements in order not to affect accuracy due to oscillator pull.

The coupling line between the rf bridge and the driven element should be as short as possible, not more than six inches to a foot, and should be coax cable of the same impedance used to feed the quad. Also the reflector and director must be in place, and connected, for these readings. The quad should be in exactly the same condition as it will be when used, except for height above the ground.

front-to-back ratio

Raise the quad in a vertical position above ground and as high as possible. You must still be able to reach the reflector coils, or the stub, whichever is used. Next, energize the quad at the resonant frequency. Point the quad away from where the measurements will be taken. This may be either a cooperating amateur a few blocks away, or your own receiver a few blocks away, or a field strength meter a few wavelengths away.

Now, adjust the stub or the coils (number of turns) until a minimum reading is obtained on the receiver or field strength meter. Measure the reflector frequency again, and if it isn't within 3 to 5 percent range there is something wrong with your measurements or your method of measuring.

standing wave ratio

In all cases, when using an swr bridge, first set the bridge "wide open," or at its most sensitive point. Then feed in just enough rf to get a full scale reading in the forward direction. Next, reverse the bridge, and read the swr. It's not necessary to hoist the quad in place to make these measurements. Quads have no open ends and thus are quite insensitive to nearby objects.

If the swr bridges were perfect, it wouldn't matter whether the bridge operated at the full-power output of the transmitter or at very much reduced power. Measuring instruments are not perfect of course, hence their limitations must be taken into account. Remember, too, that when the swr bridge is set to read reflected power, it also reads a small fraction of the incident power, and any stray coupling will affect the accuracy of the reading. Also note that the true swr will always be lower, but never higher, than the instrument indicates.

A theoretically perfect swr bridge would read the same regardless of the amount of

power fed to it. In a way, therefore, the difference between the full-power reading and the reduced-power reading is somewhat of a test for the quality of the bridge assuming that no power is leaking to the reverse diode.

Now, let's assume that the actual mismatch is 2:1 between the feed line and the antenna. This is the **maximum** you can read on the bridge regardless of where you insert the bridge into the line. If there is in fact a mismatch, you'll obtain a different reading everywhere along the line. If the true input impedance is 50 ohms, then if you measure one-half wavelength from the input point, taking the coax propagation factor into consideration, the measured impedance will be 50 ohms.

At one-quarter wavelength from the input point the measurement would be either 25 or 100 ohms, depending upon whether or not the quad input impedance was higher or lower than that of the cable. However, at exactly one-half wavelength from the feed point, it would be immaterial what cable impedance were used as long as the propagation factor is accounted for in measuring the one-half wavelength.

The reason for this is that the reflected voltage, traveling back along the line from load to source, meets the incident voltage. If these voltages meet at multiples of one-half electrical wavelength on the line, the voltages will add, because at that instant they will be exactly in phase. It is possible, with certain line lengths, to get a reading of 1:1 when the actual swr is considerably greater than this. The only way to be sure is to use several different line lengths of the same cable, and if the swr comes out approximately the same, then you're reasonably sure that the swr is being read correctly. With a perfect match the swr will be 1:1 regardless of where the meter is placed.

optimum element spacing

According to reference 1, the optimum gain occurs in a quad with 0.125-wavelength spacing between driven element and reflector. If the spacing is increased to 0.25 wavelength the gain falls off about one dB. There is an infinite number of spacings that may be used on any beam, but I found that 0.125 wavelength between reflector and driven element is indeed optimum. (The input impedance is about 50 ohms as well.) I also found that a 0.1-wavelength spacing between driven element and director, or directors, leaves little to be desired.

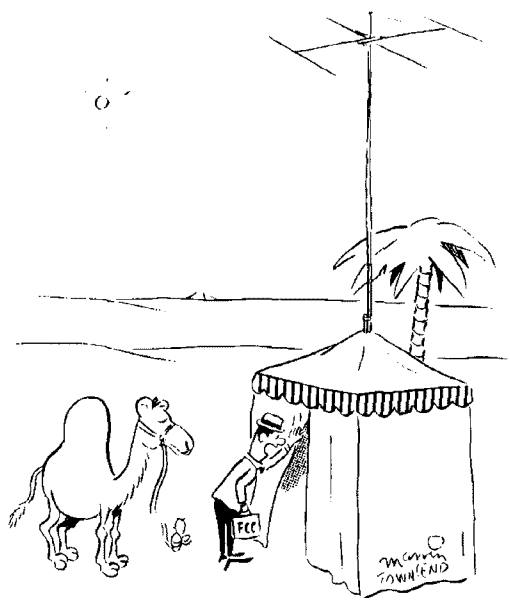
It's true that many successful quad users space the elements at considerably greater distances with possibly some advantages. But from a practical standpoint, spacings as indicated above have many advantages also, such as a reasonable size structure that can be rotated with a rotator of moderate strength, and one that can be put on a tower of the ordinary ham variety.

To obtain the excellent performance of a multielement quad (unless you want to use a boom of 25 to 50 feet), spacings of the order of 0.125 wavelength between driven element and reflector, and 0.1 wavelength between driven element and directors, are quite in order.[2,3]

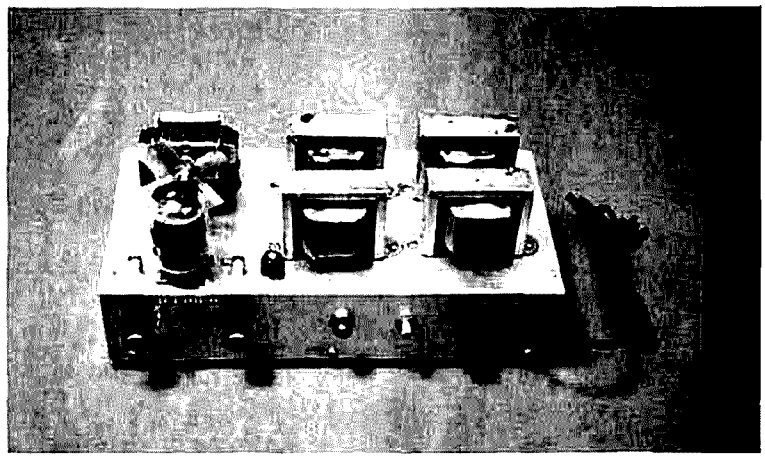
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1. William I. Orr, W6SAI, "Quad Antennas," Radio Publications, Inc., Wilton, Connecticut, 1959.
2. J. E. Lindsay, Jr., WØHTH, "Quads and Yagis," *QST*, *May, 1968, p. 11.*
3. L. W. VanSlyck, W4YM, "How Come?" *73*, February, 1968, p. 14.

ham radio



Sir! Do you realize that you are interfering with TV reception in this area?



novel linear for two meters

inductive-divider input and toroid output circuits for high efficiency

Ralph W. Campbell, W4KAE, 316 Mariemont Drive, Lexington, Kentucky

Probably the easiest way to operate on two meters is to buy a "Two'er," select a crystal for the emergency frequency, then hope someone in the nearest large city will hear you. Well, with a good beam you might be heard; then again, if band conditions are poor —forget it.

I tried to buy a linear for my Two'er that would up the output at least ten times, i.e., to 30 watts nominal. None was available that would preserve the audio quality Heath is known for and still provide a respectable signal in the nearest town (which is some 83 miles way) with anybody on two.

Obviously, to get the most from the audio in the Two'er, AB_1 operation is a must. So I built such a linear using a 7984 pentode with series-tuned input, ferrite choke loading instead of resistor swamping and a toroidal-tank output. It gave the needed punch, and now I'm heard every day even with 20-dB tropo fadeouts, if the guys on the other end are using at least one Nuvistor ahead of their receivers. The circuit described is a refinement of an earlier amplifier using a novel input system that puts the power where it's needed.

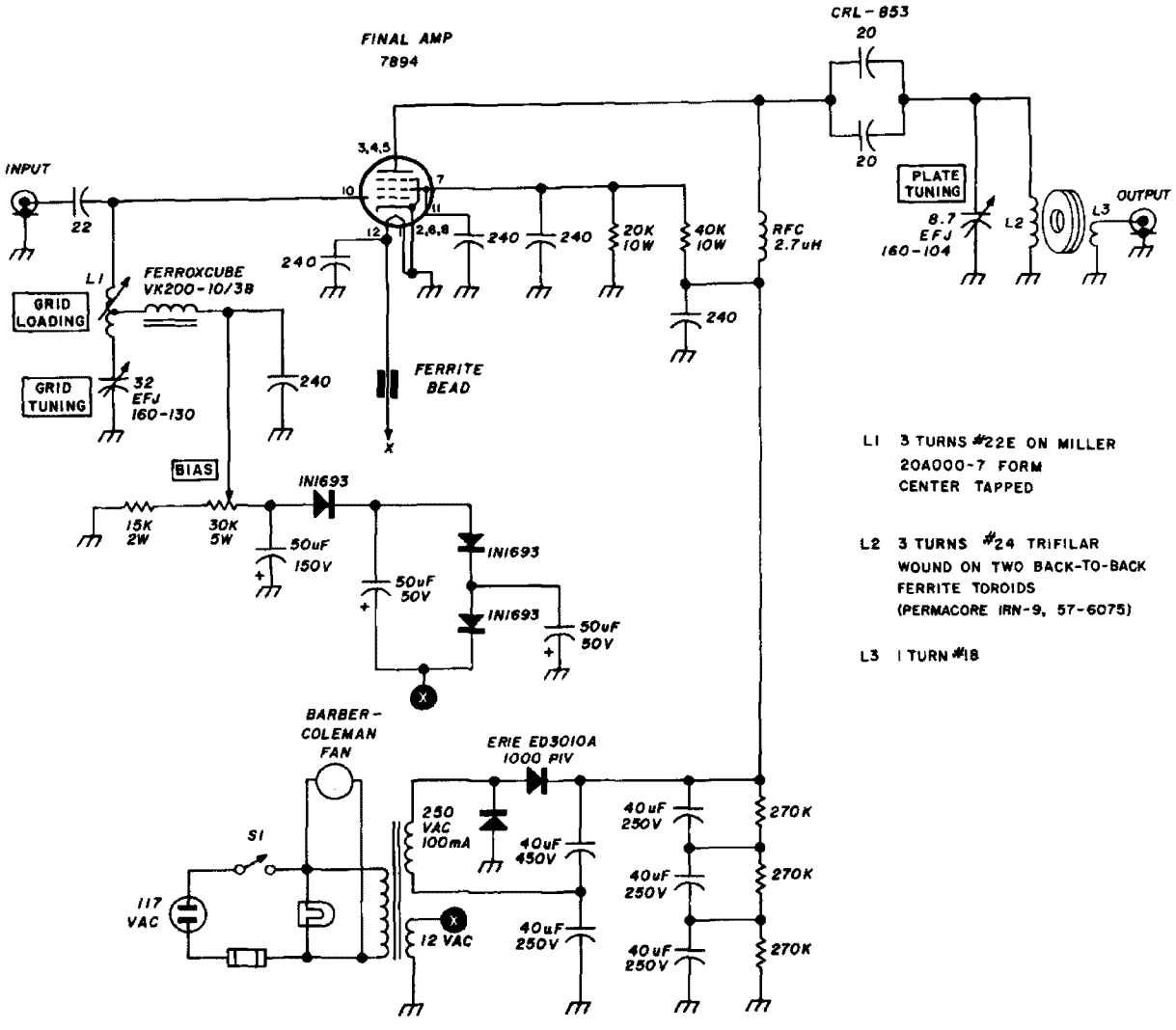


fig. 1. Schematic of novel two-meter linear amplifier. Although only one power transformer is shown here, the author used several small transformers to obtain the required voltages and currents.

development

Some of the problems I encountered involved the old TVI hangup. Living in a deep vhf TV fringe area, it's impossible to operate even class B on two meters. An exception might be a rig with a class-B single-balanced modulator with very little carrier—but even then some TVI would be a problem. In my case, however, uhf TV reception in Lexington is so popular that a dsb rig with a kW PEP final is feasible, but only because people hereabouts switch from vhf fringe channels to local uhf TV. Of course, anything like class C is "verboten."

I had a surplus of 125-V, 50-mA power transformers and several E. F. Johnson variable "M" capacitors. Most of the hardware was scrounged from other projects that never got off the runway. This included two UG-1094/U BNC connectors, 1000 PIV rectifiers rated at 750 mA and a Bud CU-622 Convertabox for the chassis.

Having built the linear "à la haywire" six months before, I tossed out the phenolic Compactron socket for a new mica-filled one, doubled the number of transformers, and used a newer Convertabox. Everything at last fit well, and most of the hole centers were eyeballed-in using a steel scriber and fierce concentration on a 10c plastic pocket ruler. Everyone around Lexington said my old 7984 bomb was loud, but the newer version was even better, and no one at either time could hear **and** see me!

Except for the parallel-tuned toroid output, over-all design of the two-meter linear is

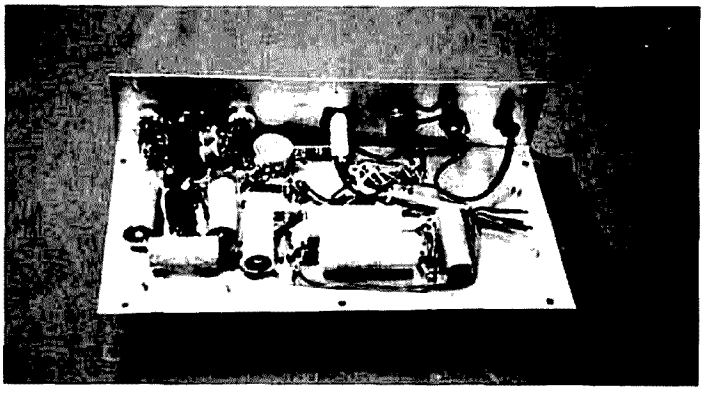
straightforward, and the series input circuit is conventional except as noted below. If desired, ceramic coil forms could be used and a physically "linear" quarter-wave line inserted in place of the IRN-9 toroids; but toroids keep the circuits small, and more power is coupled into the coaxial output and not into the wiring.

I tried a high-Q quarter-wave line and got less out. There is a limit, though, to the use of toroids on two meters. After trying unsuccessfully to get more than 25 watts measured output from a high-power PL-177WA amplifier, I have bowed a bit to the conservative by observing that a practical limit (tube dissipation basis) is 35 watts for ICAS operating conditions. This is all that seems worthwhile for AB_1 with the 7894. I did heat up some larger cores quite a bit, using the PL-177WA, but this tube is 40 watts "larger."

A new circuit feature is claimed here. As you read on you will notice that a grid-loading control is provided, saving lost watts in the output by peaking the control grid without a swamping resistor. This is really the most important part of what I have to say under "development," because by removing that swamping resistor and loading the grid only lightly you get very high efficiency in AB_1 and drive high enough to push the 7984 to its limit without an insertion amplifier.

In fact, before I accidently found that series tuning would give an rf voltage peak of the correct impedance at the control grid, I thought an amplifier would be necessary. Not so! But don't forget the trick is to center-tap the series inductance and load with a Ferroxcube VK-200-10/3B semiconductor low-z ferrite choke.*

Under chassis construction; chassis is a Bud Convorta-box.

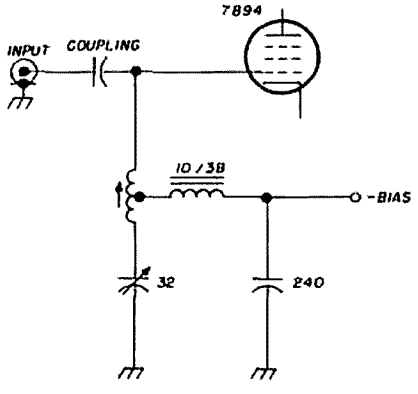


layout

Looking at the photo, from left to right we see the input connector, 30M11 input capacitor and the slug-tuned grid-loading inductance coil form. Next is the blower and 7984. Just to the right is the output BNC connector, output tuning and bias pot. The rest of the controls are as labeled.

The underchassis view shows the input tuning control; rear, Ferroxcube choke and series inductance, shield partition and heavy-duty ceramic coupling capacitors (the only kind that would handle the circulating current) and trifilar output tank circuit wound on two Permacor 3/8-inch IRN-9 sample toroids.** The rf components occupy only a small fraction of the chassis space. The rest of the chassis is used for the voltage-doubling power supply and bias-multiplier filters.

fig. 2. Input circuit used in the linear.



A series-limited neon lamp is used as the pilot and is held with a large Tinnerman speednut. The on/off switch is special in that it mounts in a 3/8-inch hole and not in the usual 7/16-inch or larger hole, which saves buying a punch for just this size.

* Order from Mr. R. Worban, Ferroxcube Corporation of America, 360 E. North Avenue, Northlake, Illinois 60164. 11c each plus shipping.

**Permacor Division of Radio Cores, 9540 South Tulley Avenue, Oak Lawn, Illinois 60453; Number 57-6075 IRN-Q material.

Button bypassing is used on both screen leads and also for plate circuit decoupling. These 240-pF buttons were salvaged from a surplus ARC-3. The plate choke is 2.7 μH, chosen for a cutoff of 149 MHz, with preference for the higher reactance obtained using the Miller chokes as opposed to the Ferroxcube. Remember that the VK-200 series are lossy when used with vacuum tube circuits for other than this application in swamping the grid. Their use in transistor circuits is a notable exception, where impedances are very low. Twenty pF was the value chosen for plate coupling, since two of these will have a series-resonant mode near the two-meter band.

Incidentally, I tried to use "dog-bone" capacitors and disc ceramics with no luck. They all burned up. When I switched from these to the more expensive ceramics I used pins 3 and 4 of the 7984 because they carry the bulk of the rf load. Pin 5 is fine for dc coupling.

the circuit

Fig. 1 shows the voltage-doubler power supply. Screen regulation is considered a myth for AB_1 operation using a pentode of such excellent structure for rf. If the plate voltage should get out of hand, so will the screen voltage, tending to reduce the plate current; also there's a bleeder and voltage divider using a 20k/10W resistor from screen to ground.

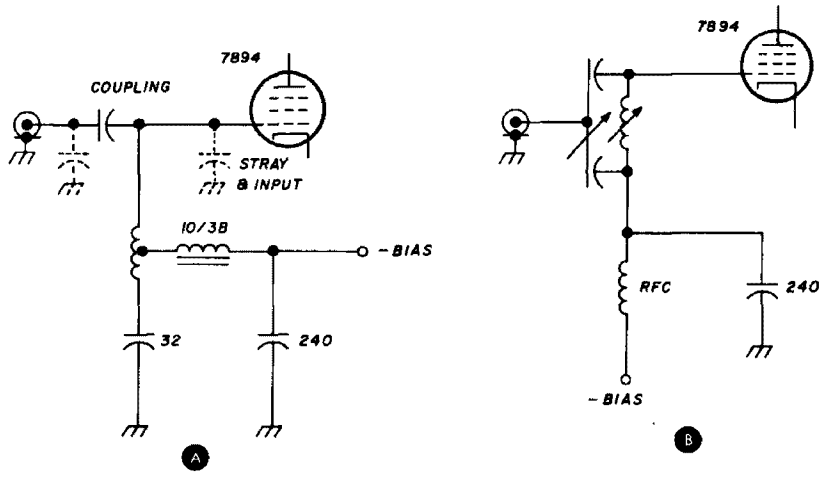
The input circuit is capacity coupled using a value equal to the input capacity of the 7984 plus strays. The ferrite choke takes the place of, say, a 220-ohm resistor and is more efficient when placed at the center of the series coil than at the grid proper. By using this approach a pole, or impedance maximum, is present with the tuning peak, and coupling is optimized by varying the inductor slug. Reactance is tuned out by the variable capacitor.

Fig. 2 is a drawing of the input circuit referred to above. Series tuning by the M capacitor is like moving up and down the VK 200-10/3B choke (made of the popular "bead" material) from center position as resonance is achieved. That is, inductive division, like capacitive division, (to quote RCA[1]) feeds the right voltage and current at two-meter rf frequencies to match the input admittance and susceptance of the high-C 7984. **Fig. 3** shows both tubes. Capacitive effects are represented by C_{in} (**fig. 3a**), with bead choke resistance swamping.

If you'll also look further into the RCA Transmitting Tube Manual, you'll notice they use only center-tapped series-tuned circuits for vhf since strays have already limited that fraction of the input which can be capacitively divided, and varying the slug of the coil form is the best way to optimize an input circuit like this.

Other arrangements are shown for high-C input vhf tubes (**fig. 3B**). In general, use at least one and one-half times the input capacity of the tube for the series variable, and

fig. 3. Different input circuits discussed in the text: inductive division in A, capacitive division in B.



always center tap the inductance. Also, use permeable cores in the form instead of the brass slug type, unless extreme bandwidth, lowered Q and extra power are available. Use of solenoid-wound rf chokes for class-C and -B operation is recommended, but use ferrite for linears that don't draw grid current.

operation

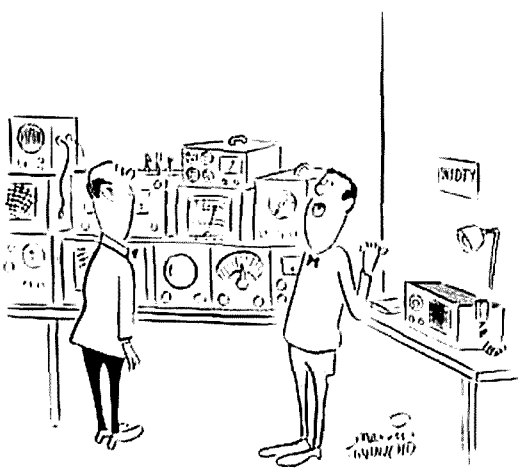
I used this linear for over a year without a blower, and have never had difficulty with it. It has a measured output of 12 watts (with a new 7984) in AB_1, and will go to AB_2 if the bias is set too low. (You can tell by listening for distortion in the monitor.) For AB_1 as opposed to class C, where the input power is specified, I chose plate dissipation as a more accurate gauge for rating, times the book value for theoretical efficiency, i.e., 33% of a maximum rating of 35 watts dissipation, which would be about 12 watts output.

On the air I say "50 watts," and this can be measured as input in the usual manner. My goal is 2000 watts dc input to two 4CX1000's in AB_1. Though of questionable legality, this would give no more than 660 watts output, like the class-C boys, with virtually no TVI.

references

1. *"RCA Receiving Tube Manual,"* 1966 edition, page 54.

ham radio



"My transceiver is over here. That's just my losting equipment."

NOISE BLANKER
FOR THE SWAN 250

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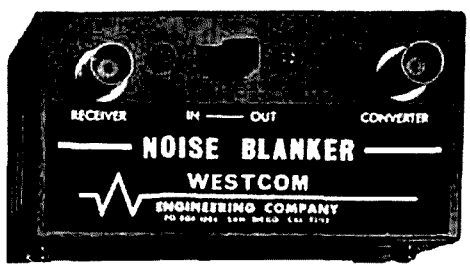
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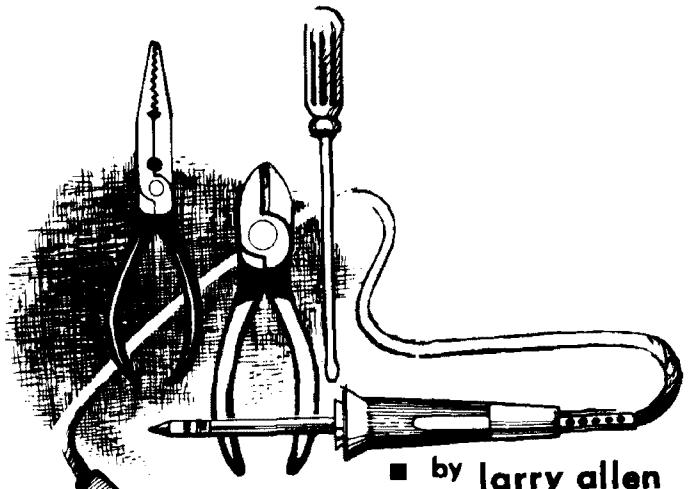
Refer to the New Products column of the August '68 issue of Ham Radio Magazine for additional information on the TNB Noise Blanker or write for technical brochure.

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by larry allen

the repair bench

more ohmmeter troubleshooting

Last month, I explained how to use your ohmmeter for safe troubleshooting. Before I ran out of space, I reviewed series and parallel paths and also told you how to get accurate readings on your ohmmeter.

I hope you learned to recognize the "odd" series-parallel paths I described. When you're tracking down a fault in some receiver or transmitter, how easy it is for you depends on how well you spot the resistance paths that aren't always obvious. This month, I'll start off by telling where you might find some of those paths.

hidden resistance paths

There are two small parts most likely to cause false ohmmeter readings: electrolytic capacitors and transistors. Electrolytics are part of all power supplies and are used for decoupling along dc supply lines. In either use, they often get in the way of ohmmeter tests. Transistors foul up ohmmeter readings because of their "diode" nature. To make matters worse in transistor equipment, many coupling capacitors are electrolytic. They mix their leakage currents with the transistor leakage currents, and you can drive yourself nuts sorting out the true paths.

Fortunately, these two paths follow known behavior patterns. If you know the patterns, you can figure out if you're being misled. Here are some of them.

1. In **power-supply paths.** The leakage resistance of most electrolytic capacitors is nearly 50k. Learn to allow for it when you're checking resistances along B-plus lines. The diagram in **fig. 1** shows power-supply paths in one receiver schematic (simplified, of course).

Suppose you clipped the common lead of your ohmmeter to ground and the test lead to the plate of V4. Since there is no apparent path to ground, you might expect to read an open circuit, or infinity. Instead, you read around 600k. As you can guess, the ohmmeter is reading the resistance of R13 **plus** the leakage resistance of the power-supply capacitors.

2. Allow for **diode action.** Suppose the reading from the plate of V4 to ground is almost exactly 470k. What does that mean? Maybe a short in one of the electrolytics? To check, you move the test lead to the junction of R3-C4. A very low resistance there seems to mean there's a shorted electrolytic.

But before you jump to a false conclusion, reverse the test leads: put the common lead on the R3-C4 junction and the other (red) lead to ground. The low reading disappears!

It's confusing, isn't it?

What happens is that the voltage of the ohmmeter battery causes the rectifier diodes to conduct and give the low reading. Whenever the negative side of the ohmmeter battery (sometimes, but not always, the red test lead) is touched to the cathode end of the rectifier, the ohmmeter reads the diode's forward resistance (which is low). An ohmmeter thus connected from R3-C4 to ground might read through R1-R2-R3 and the diode. Connected to the plate of V4, the same ohmmeter might read 470k almost exactly, since the 1k (approximate) resistance of the R1-R2-R3-diode path is tiny by comparison. In either case, reversing the test leads eliminates the false path.

3. Consider **transistor leakage.** A form of diode action takes place in transistors. That is, the ohmmeter battery can forward-bias a transistor junction and make it draw current. The ohmmeter then reads the junction resistance as well as whatever circuit-path resistance you're measuring. There's an example in **fig. 2.**

The pnp transistor is wired up as a mixer. Suppose you connect your ohmmeter across R1 to measure it. Instead of the expected 4.7k, you read about 1k and figure R1 is bad. However, if you reverse the ohmmeter leads, you measure 4.7k. How come?

The base-emitter junction of a pnp transistor is forward-biased when the base is negative and the emitter is positive. If you connect the test leads the wrong way, the ohmmeter reading includes the base-emitter

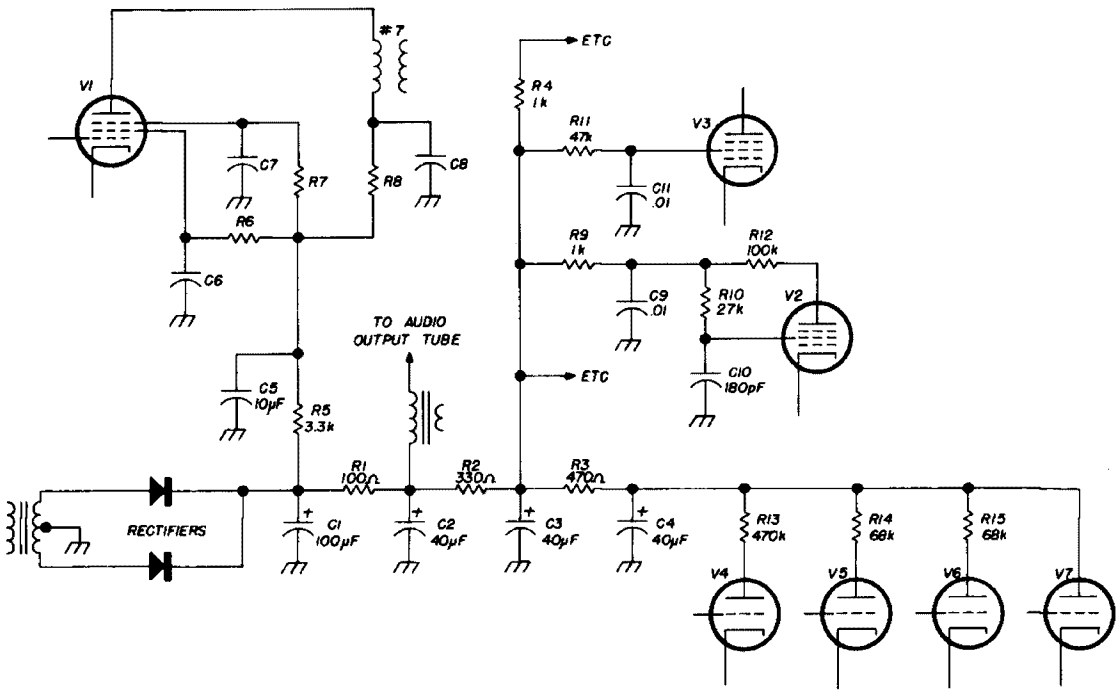


fig. 1. Power-supply from a popular receiver. It's simplified, to make tracing easy. If you must troubleshoot such a set, you can draw a simplified diagram like this to help you "see" the circuit branches.

resistance (100 ohms or so). It is in series with R3, and both are in parallel with R1. The result: a wrong measurement.

How about when you measure R2? There should be no diode action between base and collector. What would you think, then, if you tried measuring R2 and got a 40k reading? Another hidden path is probably the cause. **Fig. 2B** shows the possible cause: leakage through the power-supply electrolytic capacitor (which may not even be drawn at that location on the schematic). The path through R4, the capacitor's leakage, and R1 lowers the resistance read by an ohmmeter connected across R2. You can thus get an

erroneous reading, even if the transistor is normal.

A significant clue, though it might fool you, is the fact that reversing the ohmmeter leads in this case doesn't change the reading. The trouble is therefore **not** diode action.

4. Watch out for **paths drawn elsewhere.** The path just described is a common example. There are others harder to spot. One is shown in **fig. 3.**

The situation in **fig. 3A** is apt to be overlooked in tube circuits. The trouble is that schematics show only the common power-supply connections. If there happens to be a ground path somewhere else on the schematic, you can get a low ohmmeter reading you can't account for.

In **fig. 3A,** for example, if you measure from the plate of VI to ground, the voltage-divider network in the screen circuit of V9—which may be drawn at the other end of the diagram—causes an incorrect reading. (It should be infinity.) Even if you disconnect the power-supply capacitor, the too-low reading persists. The equivalent circuit is redrawn in **fig. 3B** to show how the circuit looks to the ohmmeter. (R9 has no effect, since it does not go to ground.)

You'll run into this voltage-divider situation more often in transistor sets than in tube equipment. Base-bias and emitter-bias dividers are standard practice. They cause trouble only when they are in parallel with each other but are widely separated on the schematic.

5. Something else to watch for in transistor circuits is shown in **fig. 4.** This one can fool you for several reasons. First, there is the diode action of the base-emitter junction. Second, there are parallel paths created by the power-supply electrolytic (not shown, as usual). Third, the parallel path depends on where you connect the ohmmeter.

The four rearrangements of the circuit show what the ohmmeter "sees" as it is connected between ground and each of the four points: A, B, C, and D. Anytime you have trouble figuring out paths, redraw them the way I have these.

From A, the ohmmeter should measure either about 850 ohms or about 4.6k. The diode is the reason for the alternative. One way, the ohmmeter makes the diode conduct and measures its forward resistance plus R4; the other way, it reads only the resistance of R1, since the diode can't conduct. The resistances of R1, R2, and R3 are comparatively so high, they have little effect even though they are in parallel.

From B, the ohmmeter should measure about 35k. The diode shows little effect, because R1 and R4 are relatively small compared to R2. The ohmmeter therefore reads about the same either way. The 68k–72k is in parallel with the (approximate) 77k.

From C, there are several considerations. First of all, you don't have to worry about the effects of R2, R3, and the power-supply electrolytic. They are so much larger than R1 that the 4.7k is the only resistance that can affect that leg of the circuit. The diode action of the base-emitter junction is important, though. Connected one way, the ohmmeter

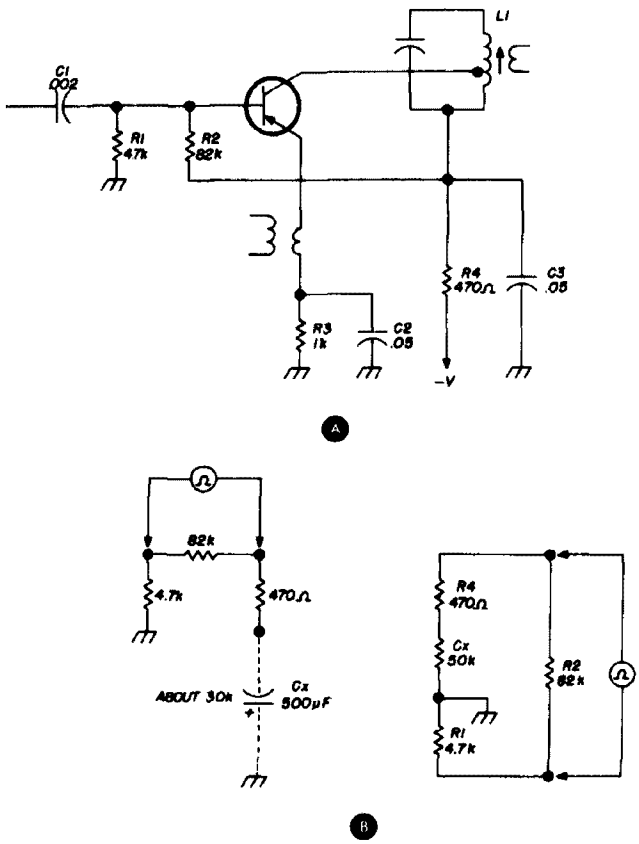
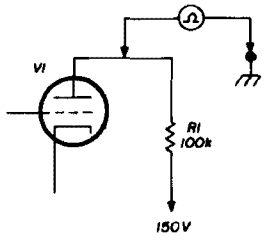


fig. 2. Hidden parallel paths in a transistor circuit.

reads only the 820 ohms of R4. Connected the other, it reads slightly less because diode action puts R1 in parallel; around 750 ohms seems about right.

From D, the ohmmeter mainly reads the value of leakage in the power-supply electrolytic—about 30k. The much higher values of R2 and R3 isolate any action of the base-emitter junction, or the effects of R4 or R1. The condition of the electrolytic determines exactly what the ohmmeter reads.

fig. 3. Sometimes a parallel path is hidden by the way the schematic is drawn.



knowing the cause of wrong readings

These general descriptions of where to look for "hidden" parallel paths also suggest ways to deal with them. When your ohmmeter readings are different from what the schematic leads you to expect, here are some quick steps to run down the cause.

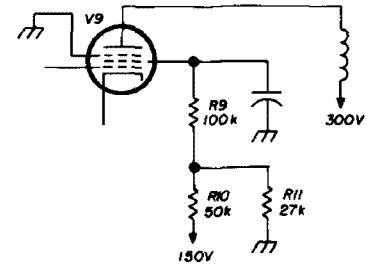
1. If a reading is low, reverse the ohmmeter leads. If the reading increases, there's diode action in the circuit, which may be caused by a diode, a power-supply rectifier, or a transistor. In unusual cases, **slight** differences may be traced to electrolytic capacitors that don't "form" under reverse polarity.

2. If an ohmmeter reading starts low and builds up slowly, an electrolytic capacitor is causing it. This is the normal charging action of the electrolytic. Reversing the ohmmeter leads gives a lower reading, which builds back up again to the higher one. The reading is anywhere from 25k to 50k, depending on the condition of the electrolyte in the capacitor. The charging action is what you watch for.

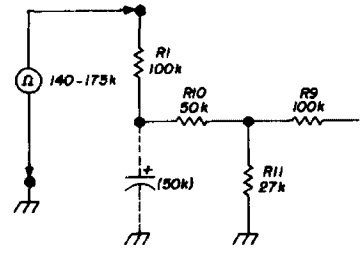
3. If you're still looking for the reason for a low reading, see if you can find any other parallel paths drawn elsewhere on the schematic diagram. If you don't find any, you are probably on the trail of the fault.

what to do about parallel paths

There are times when it's faster and easier to eliminate the effects of parallel paths or to get around their effects some way. Once you spot a wrong reading, track down the guilty parallel component. You can't always figure it out from Ohm's law, either.



A



B

One of my **first** suggestions is: Don't make resistance measurements with respect to ground. Don't clip the common lead of your ohmmeter to ground. Make each measurement between specific points **in** the circuit.

As an example, think about a circuit like the one in **fig. 1.** You can clip the common ohmmeter lead at the junction of R2-R3, and measure the branches through R9, R11, R4, etc., directly. That way, you eliminate any effects from power-supply electrolytics. To measure in the branch that starts through R5, you can clip the common lead at the junction of R5-R1 and eliminate any diode-action problem with the rectifiers.

And so on. By making point-to-point measurements, you don't have nearly so many parallel paths to consider. Also, as you move the main test lead on out a branch, it's easier to figure out the effects of added circuit resistances. Analysis is much simpler.

My **second** suggestion is to disconnect unwanted parallel paths when they get in your way. This happens seldom, if you plan your point-to-point measurements. But when it does happen, it's usually simple to unsolder

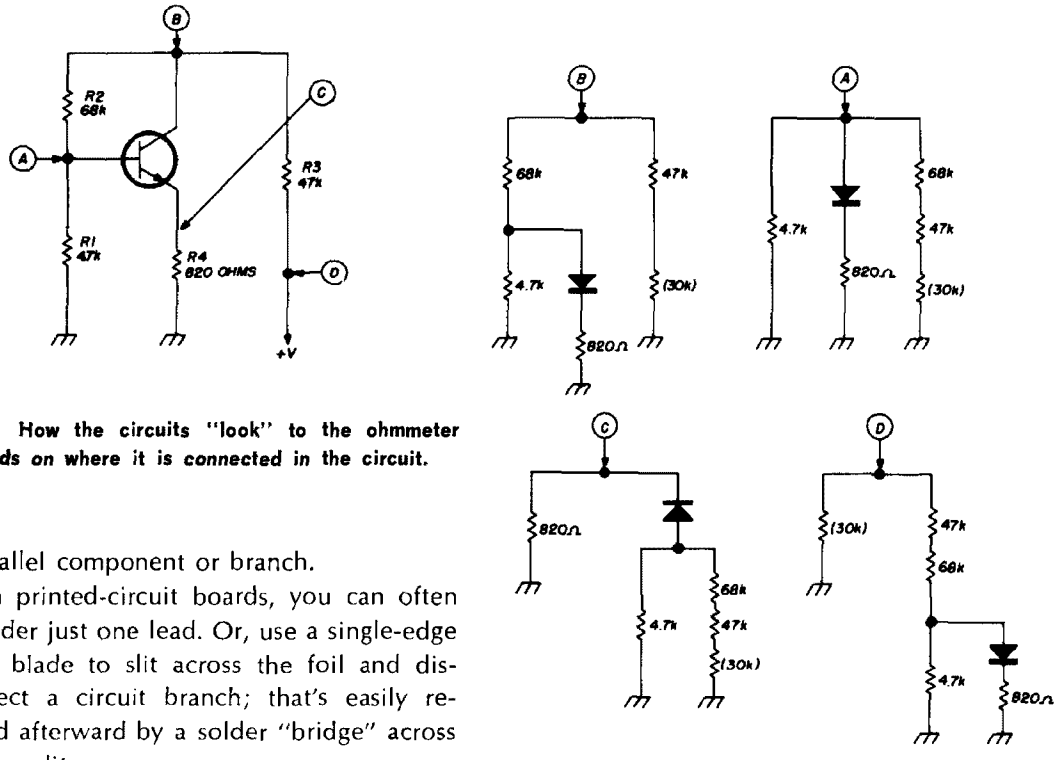


fig. 4. How the circuits "look" to the ohmmeter depends on where it is connected in the circuit.

a parallel component or branch.

On printed-circuit boards, you can often unsolder just one lead. Or, use a single-edge razor blade to slit across the foil and disconnect a circuit branch; that's easily restored afterward by a solder "bridge" across the tiny slit.

My **third** suggestion is to pull a tube or transistor out of its socket if you think it's fouling up your attempts to interpret resistance readings. Transistors that are soldered into their printed-circuit boards aren't that simple, yet can be handled without much trouble. Just unsolder the base connection (or slit the foil as already described).

In a circuit that contains a FET (field-effect transistor), unsolder the gate connection **and** the source connection. You can get low-resistance readings through some FET channels, even when the gate is disconnected.

My **fourth** suggestion for dealing with parallel paths is a last resort. When you can't analyze a circuit any other way, disconnect and test each part in the circuit—one at a time. It takes time, but if you've done your preliminary testing right, there aren't too many parts to test. At least you'll have narrowed the trouble down to one small section of the circuit.

in dangerous circuits . . . the safe way

I brought up resistance measurements in the first place to explain how to trouble-shoot transmitters without exposing yourself to dangerous high voltages. Hunting a fault in the "live" plate circuit of an ssb linear can be a real pain in the neck. You need a special meter to measure high voltage, and usually you can't even reach test points without cheating a bunch of interlocks. There may be an overcurrent relay, too, or an undercurrent relay. In other words, the practical way to troubleshoot is with the transmitter or linear completely disconnected from power, and opened up. For that, you need ohmmeter measurements.

The plate circuit of a commercial linear

amplifier is shown in **fig. 5.** Mainly, you check for continuity. However, you can disconnect the power-supply lead and make measurements between ground and points in the circuit.

If an interlock switch isn't working, your measurements show it. The normally open (n.o.) switch is there to short out high voltage across the capacitors when you open the transmitter compartment. At the same time, the normally closed (n.c.) interlock switch removes applied voltage. Start your tests with power disconnected and the interlocks pushed in the way they would be with the compartment closed.

If everything's normal, you should get a low resistance reading from the plate cap of either tube to the clip that connected to the power-supply terminal. Check the opening and closing of the overcurrent relay with your finger. An open choke, tuning tank coil, relay contact, relay coil, n.c. interlock, or metering resistor (R1) spoils continuity, but each one is easy to track down. Just clip your common ohmmeter lead to the power-supply clip, and work your way to the tube plate-caps with the other test lead.

A shorted or leaky by-pass capacitor (C1, C2, C4) shows up when you connect your ohmmeter, set for its highest megohms range, between the plate-cap and ground. Be sure the PA current meter is not connected; it could be an unseen parallel path. Disconnect the power-supply clip, too, for the same reason. Set the ohmmeter to its lowest range to test the n.o. interlock switch; it should read zero ohms to ground when the cover is off the transmitter high-voltage compartment and infinity with the cover in place.

In high-voltage circuits, especially the ones that carry high-energy rf, certain troubles only show up when the circuit is energized. For example, capacitor C2 or C1 might break down under high-voltage stress, yet test okay by the ohmmeter. The only sure test is to remove it from the circuit and substitute another one. This is costly, though, when the capacitor is an expensive one, such as a bathtub type. For these unusual cases, there are some special tests I'll talk about in a future column. You can make these tests **without** working inside the "live" transmitter or linear amp.

individual parts tests

As a wrapup for this two-month coverage of ohmmeter tests, here are some parts tests. Usually, these tests are made with the part disconnected from its circuit. When other tests and resistance readings point to a particular part, the final test is of the part itself. The following tests are quick and dependable.

Test **resistors** and potentiometers by measuring their resistance. For potentiometers, also check how smoothly the resistance changes. Connect your ohmmeter between one end lug and the center lug, and turn the control shaft slowly; any unusual or sudden resistance change means you need a new control.

Test ordinary **capacitors** by measuring with the highest range on your ohmmeter. There should be absolutely no leakage (infinite resistance). Touch the two capacitor leads together for a moment before applying the ohmmeter. Capacitors above 0.01-μF show a slight charging "kick" of the ohmmeter pointer. Those that don't are probably open. A shorted capacitor reads zero ohms between the two leads.

Test **electrolytic capacitors** by applying the ohmmeter leads in first one direction and then the other. Settle for the connection that gives the highest reading. It should exceed 30k, but not go beyond 60k or so. If it's too low, the capacitor is leaky. If it's too high, the capacitor is becoming ineffective. Notice how long it takes to "build up" to the highest ohmmeter reading. If it takes more than 3 or 4 seconds, the capacitor needs replacing.

There are three tests for **coils and transformers.** Test them first for continuity of windings. The schematic diagram may list the dc resistance of each winding; if some turns are shorted, the reading will be much too low. If a winding is open, the ohmmeter shows nothing.

Test coils and transformers second for leakage between windings. Use the highest ohmmeter range. Even several megohms—an awfully high resistance—is no good between transformer windings; they should be totally isolated (infinite resistance).

Test coils and transformers third for leakage from windings to ground. With the transformer mounted as usual but with all its leads disconnected, clip one ohmmeter lead to

ground and touch the other to each winding lead. There should be infinite resistance, since each winding should be totally isolated from the frame of the coil or transformer. If you think maybe the coil frame isn't grounded, clip one ohmmeter lead to the frame and touch the other to each winding. There should still be no reading. A reading other than infinity indicates leakage.

Test a **relay** by checking its coil the same way you test a transformer winding. Check the coil for continuity; if you know the dc resistance, measure it. If the ohmmeter reads infinity, the relay coil is open. Then check the relay contacts, with the Rx1 ohmmeter range. Normally closed contacts should measure zero ohms when the relay is at rest; at the same time, normally open contacts should measure infinity. Push the armature in with your finger. The normally open contacts should measure zero ohms that way, and the normally closed contacts should measure open (infinite resistance).

Test **speakers and headsets** by measuring them with the Rx1 ohmmeter range. **Listen** as you make the ohmmeter connection; the voltage of the ohmmeter battery usually makes speakers and headsets "pop." The ohmmeter pointer should show very low resistance for speakers and a few hundred ohms for headsets.

Test **microphones** in one of three ways, depending on type. For **carbon** microphones, connect the ohmmeter across the element leads and whistle into the mike. The reading should vary with each whistle. If it stays steady, the granules are packed and the element needs replacing. If the ohmmeter reads infinity, the mike cord is probably open, or the element is making poor contact.

For **dynamic** microphones, the exact resistance depends on the impedance of the mike. However, there should be some sort of reading, even if an input transformer is used inside the mike. Also, if you listen, there is usually a "pop" in a dynamic mike when you connect or disconnect the ohmmeter.

For **ceramic** microphones, the easiest test is to connect the ohmmeter and listen for a slight "click." The ohmmeter shows a resistance reading only if the mike is defective; a good ceramic or crystal mike reads infinity resistance.

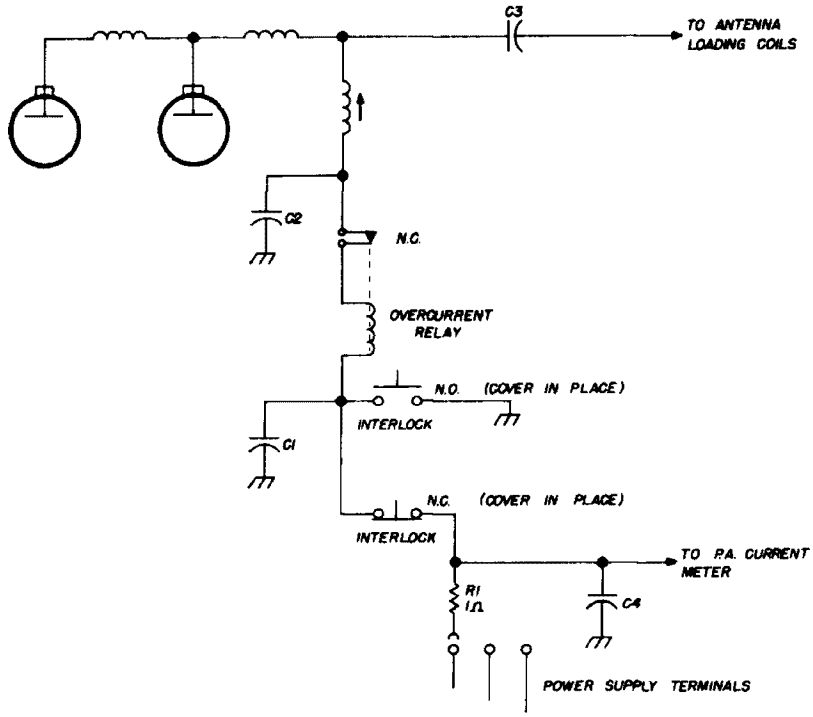


fig. 5. High-voltage circuits are where resistance troubleshooting serves best because dangerous voltages are gone.

Finally, connect the ohmmeter across the **push-to-talk** leads, and operate the button on the mike; a zero-ohms reading when the button is held means the p-t-t contacts and wires are okay.

Test **diodes** by measuring them first one way and then the other. A diode should measure low resistance in one direction and high resistance in the other; if the two readings are closer together than 100-to-1, the diode is probably bad. Power-supply rectifiers usually measure 100–200 ohms in the low-resistance direction; other types measure much lower.

A conventional **bipolar transistor** should show high resistance between emitter and collector, no matter which way the ohmmeter is connected. Between the base and emitter, however, resistance should be low in one direction and high in the other—just as in a double. If the base-emitter resistance is high in both directions, the transistor is open; if it's low in both directions, it's shorted. If resistance between emitter and collector is lower than several thousand ohms in either direction, the transistor is probably useless.

A **field-effect transistor** normally shows low resistance between drain and source—usually no matter which way the ohmmeter is connected. It should show extremely high resistance in one direction between gate and drain, and fairly high in the other direction. A MOSFET, sometimes known as an IGFET, should show infinite resistance between its gate and any other element, no matter which way the ohmmeter is connected.

next month

Everyone has his pet way to troubleshoot. In past **repair bench** columns, I've shown you several ways. You've read about signal tracing, signal injection, and now resistance measurement.

The best commercial technicians use several methods in various combinations that fit particular servicing problems. In my next column, I'll talk about one that is probably the most common: voltage testing. When a part goes bad, it usually upsets the dc operating voltages in its circuit. Those voltage changes, if you know how to interpret them, are a valuable clue to what part is bad. Next month, you'll find out how.

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propagation predictions for january

Victor Frank, WB6KAP, 12450 Skyline Boulevard, Woodside, California 94062

Late reports that reached me in October indicated that ionospheric critical frequencies were not much higher (about 10%) than predicted by ESSA. Peak values at Raratonga in the Cook Islands (latitude 21° S, longitude 160° W) rose to almost 15 MHz, compared to about 20 MHz during the peak of previous cycles. Local midnight critical frequencies have been running as high as 12 MHz, however. On the West Coast, peak values of critical frequencies have been running as high as 11 MHz, again about 10% higher than predicted.

We have reports of 50 MHz F2-layer openings during the first half of October between Hawaii and the Marshall Islands, South America, Mexico, and even (weakly) the West Coast of the United States; also between Japan and the Marshall and Cook Islands. Transequatorial scatter propagation between Hawaii and the Cook Islands was an almost nightly occurrence. However, the big, strong signal F2-layer openings into temperate latitudes hoped for this fall have so far been absent.

Quite a few six-meter operators have been missing a good bet by not running more nightime schedules with and listening for beacon stations in the southern hemisphere. I was pleasantly surprised during the latter part of September to hear ZK1AA's beacon station on 51.022 MHz three nights in a row between 11 PM and 1 AM local daylight time. He was weak, both here and in Southern California at W6NLZ, but you'd never have guessed that 51 MHz was open to the South Pacific by monitoring the low edge of the band.

the active sun

Let's continue our discussion of the sun, which last month stressed its stability. The sun rotates in the same direction as the earth. An observer at the north star would see both the sun and earth rotating in a counterclockwise direction as the earth revolves counterclockwise around the sun. Thus the solar surface rotates from our east to west. Since the axis of the sun is inclined 7° to the ecliptic, we are able to see the sun's North pole only from June to December.

The sun does not rotate as a solid body—the equatorial regions rotate faster (25-day period) than the polar regions (30-day period). It is this equatorial acceleration and the presence of a general magnetic field that is the cause of the solar cycle. The sun's general field penetrates the surface only in polar regions (greater than 55° latitude) and is called a poloidal field. The intensity of the poloidal field is about 1 gauss, compared to fields of 0.24 to 0.61 gauss observed on the surface of the earth. At solar latitudes less than 30°, the general magnetic field is confined to a sheath of about 5% of the solar radius. At the equator, the general magnetic field is believed to have an intensity of about 5 gauss.

The motions of ionized media (plasma)

are closely coupled with the magnetic and electric fields. Relative motion between a plasma and a magnetic field is inhibited since such motion would produce an electric field which would result in large current flow (since the plasma is an excellent conductor). Therefore, the plasma is forced to move along magnetic field lines; otherwise the magnetic field is dragged along with the plasma.

The differential rotation wraps the subphotospheric field lines around the sun to form a toroidal field. The toroidal field strength continually rises as the field lines

Sunspots are believed to develop in pores; a pore is a small (1000- to 2000-mile diameter) moderately dark area which may occur anywhere on the sun and has a normal lifetime of a few hours. Only pores within centers of activity develop into sunspots. The surrounding hotter plasma tends to flow into a pore, carrying its magnetic flux along and concentrating it.

As the magnetic field strength grows, convection from lower hotter levels is decreased, the surface temperature drops, and the spot grows. At full development, the spot may be 20,000 miles or more across and

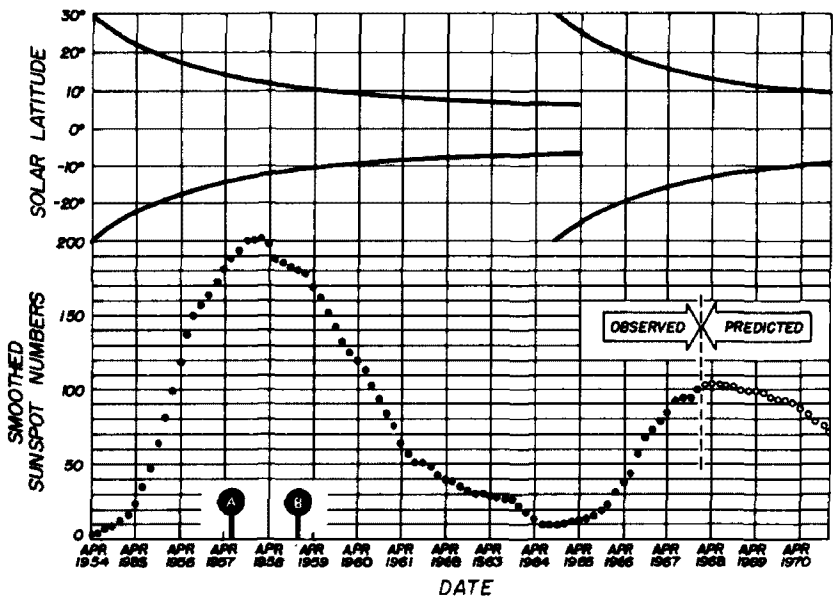


fig. 1. Smoothed sunspot numbers with latitude of maximum occurence (theoretical). The south poloidal field reversed near date A, and the north poloidal field reversed near date B.

are stretched and crowded together like a rubber band that has been wound up tightly. When the magnetic field strength reaches a critical value of about 250 gauss, tubes of magnetic flux may "bubble up" through the photosphere to form bipolar magnetic regions. It is in these "centers of activity" that sunspots, solar flares and filaments occur.

sunspots

Sunspots are visible as small dark areas on the solar disk. They appear dark because their temperature is only about 3000° K compared to 6000° K for most of the photosphere. The low temperature is due to the presence of high magnetic fields (about 3000 gauss) which inhibit convection.

consists of a dark umbra surrounded by a penumbra with very fine radial filamentary structure. It may last from as short as a few days to as long as a hundred.

The appearance and motion of sunspots, bipolar magnetic regions and other solar activity follow an eleven-year cycle, or, more properly, a twenty-two year cycle, since the magnetic fields reverse every eleven years.

Sunspots tend to occur at high solar latitudes early in the 11-year cycle and at low latitudes late in the cycle. The distribution of spot latitude vs age of the cycle follows a curve of

$$\sin \varnothing = \pm \frac{1.5}{n+3}$$

where Ø is the latitude of maximum occurrence and n is the age of the cycle in years. This theoretical curve is plotted on **fig. 1** along with the observed Zurich smoothed sunspot numbers from April 1954 to December 1967 and ESSA predicted smoothed sunspot numbers through 1970.

Over 84% of the sunspots occur in bipolar pairs. The preceding and following (with respect to solar rotation) spots have opposite magnetic polarity, and the preceding spot is closer to the equator. In the opposite hemisphere, the magnetic polarity of preceding and following spots is reversed. The preceding spot has the opposite polarity as the polar field in the same hemisphere at the beginning of the cycle. This is consistent with the theory of spot development from the twisting of the polar field due to equatorial acceleration.

The preceding spot is generally larger and has more magnetic flux (by a ratio of 3:1) than the following spot. Sunspots rotate about their axes (which are tilted in the forward direction) in the direction of the Coriolis force* (counterclockwise in the Northern Hemisphere, clockwise in the Southern Hemisphere). In general, however, the sunspots are calm magnetic lands in a sea of great turbulence that forms the rest of the bipolar magnetic region and the whole of the solar disk by comparison.

Sunspots are a secondary phenomenon since the bipolar magnetic region forms first; they themselves do not radiate increased ultraviolet and x-ray emission that affects the earth's ionosphere. In fact, the

* Usually defined as a deflecting force acting on a body in motion (as an airplane or missile) due to the earth's rotation. In this case it refers to the force on a sunspot due to the sun's rotation.

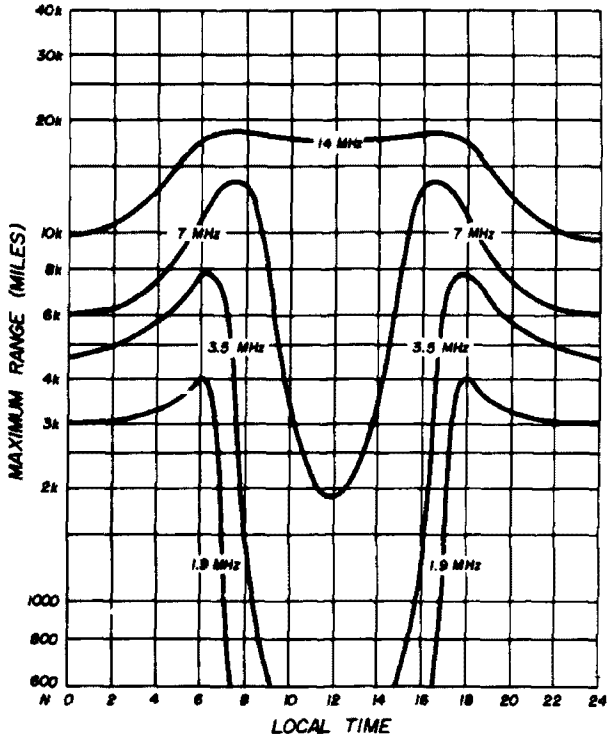


fig. 3. Maximum range to the north due to absorption.

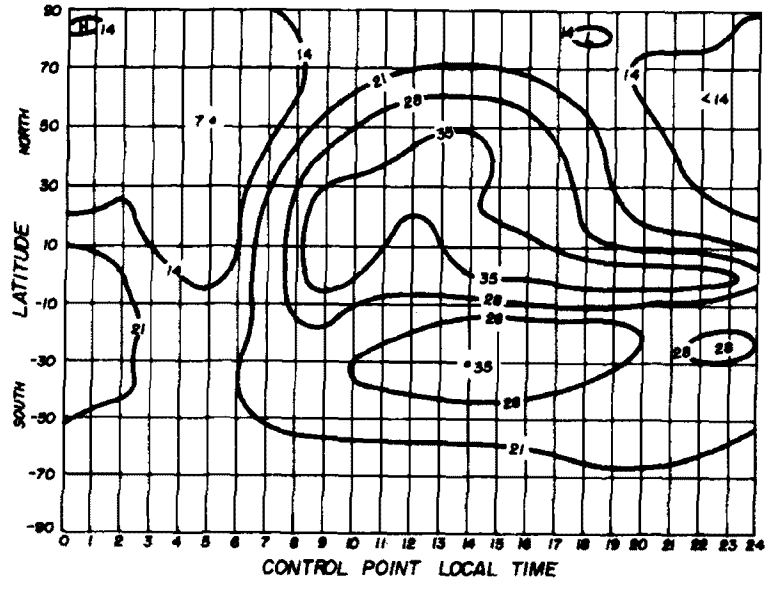


fig. 2. Time chart of maximum usable frequencies for January, 1969 centered on 90° West longitude.

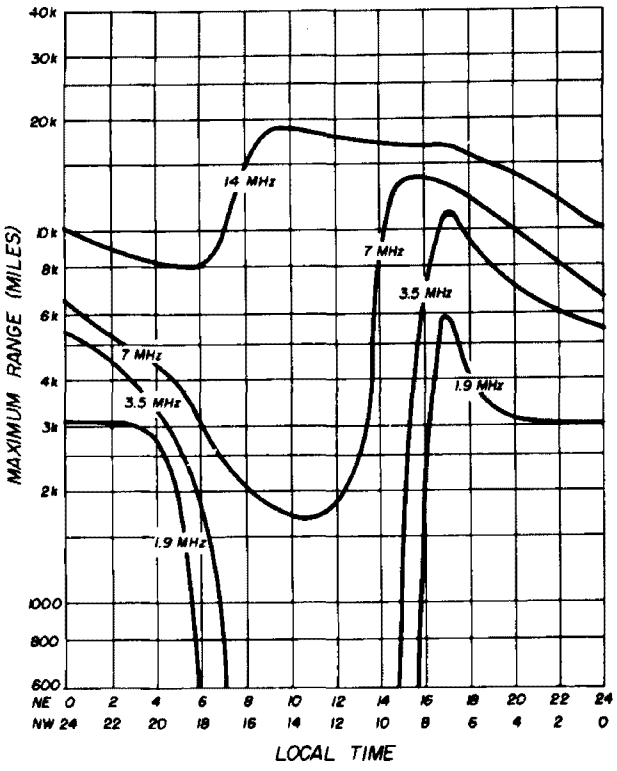


fig. 4. Maximum range to the northeast (top time scale) and to the northwest (lower time scale) due to absorption.

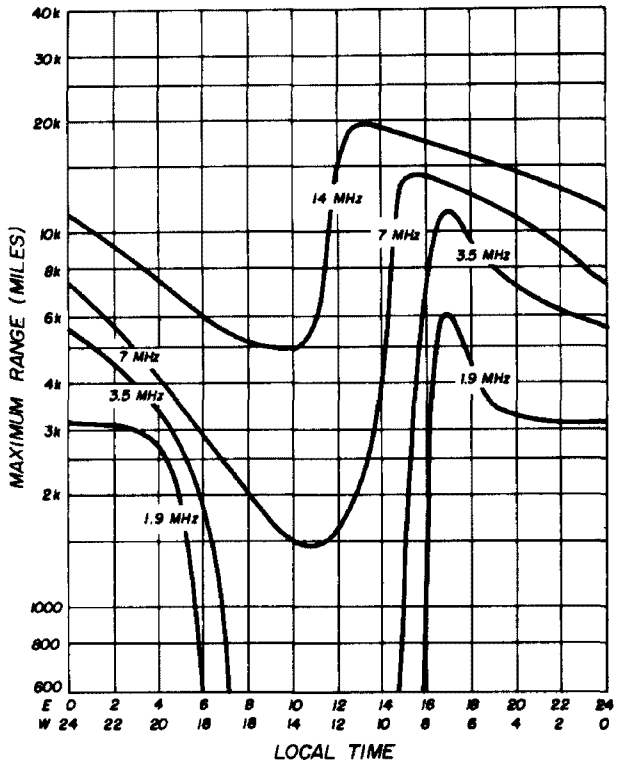


fig. 5. Maximum range to the east (top time scale) and to the west (lower time scale) due to absorption.

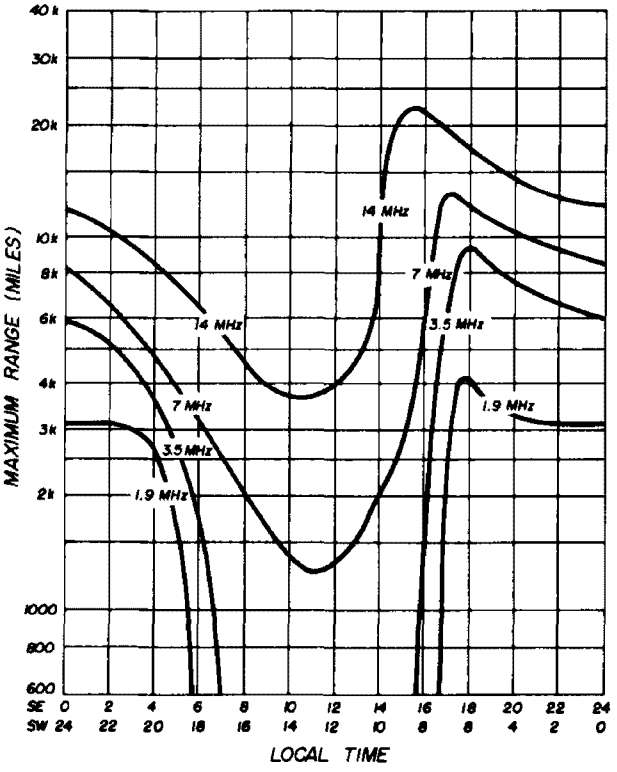


fig. 6. Maximum range to the southeast (top time scale) and to the southwest (lower time scale) due to absorption.

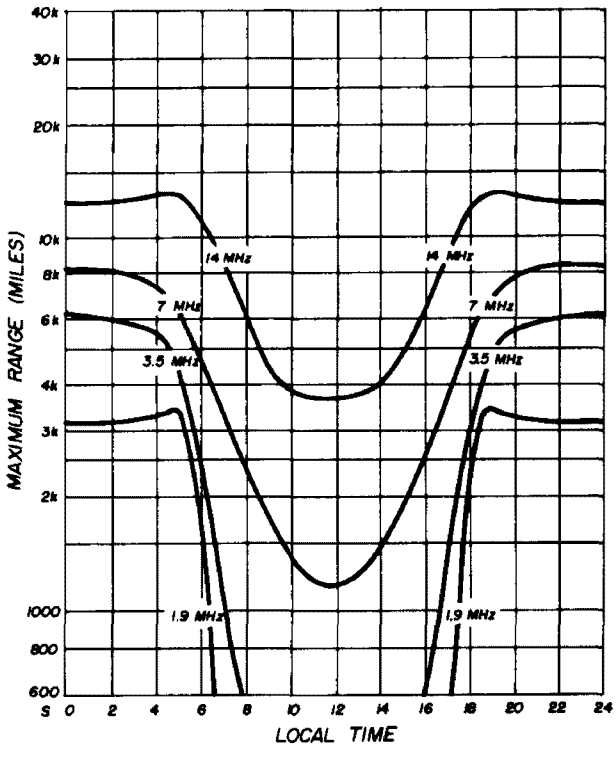


fig. 7. Maximum range to the south due to absorption.

total radiant energy output per unit area in a sunspot may be only a tenth of normal for the solar disc.

Sunspot numbers** are by no means as accurate an index of solar activity as you might wish, since they are somewhat artificial in nature. For example, a large sunspot pair somewhere on the solar disk would result in a Wolf number 12; add an insignificant small spot somewhere else on the surface, and the Wolf number becomes 23, a jump of almost a factor of two. This artificial nature of the sunspot numbers is one of the reasons for the poor correlation between monthly sunspot numbers and monthly average ionospheric critical frequencies. The answer is a better index of solar activity, such as the combined area of all the sunspots, or of bipolar magnetic regions, or (now that we have space probes) of the magnitude of the ionizing radiations that affect our ionosphere.

metamorphosis

The magnetic flux lines joining the two poles of a bipolar magnetic region may expand outwards to great distances in the corona—and may break away with the flow of plasma away from the sun. As the solar cycle progresses, the preceding regions of flux migrate towards the equator and neutralize each other; the following regions of flux migrate towards the poles where they first neutralize the old poloidal field and then from a new poloidal field of reversed polarity. The energy for maintaining and concentrating the magnetic flux comes from the kinetic energy of differential rotation.

The reversal of the poloidal field takes place near the maximum of the solar cycle, but strangely enough, both poles need not reverse simultaneously. At the peak of sunspot activity in the southern solar hemisphere in the middle of 1957, the southern poloidal field reversed, but the peak of sunspot activity and reversal of the poloidal field in the Northern Hemisphere did not occur until the end of 1958. These reversals left the poloidal field of the sun parallel to that of the earth. One or both solar poloidal fields may have already reversed this year.

** Sunspot numbers were discussed in detail in the November, 1968 issue of *ham radio*, page 72.

propagation for january

A time chart of predicted 4000-km F2-layer muf for January 1969 is given in **fig. 2.** Maximum range versus time of day charts for 160 to 20 meters are shown in **fig. 3** to **7.** The first half of January will be a repeat of December with conditions becoming more like those of November as the month progresses. It will be a great month for high-frequency propagation in the Northern Hemisphere from frequencies as low as 1.8 MHz at night to as high as 40 MHz during the day.

The 50-MHz operator aspiring to WAC is now wishing that (in addition to a kilowatt, a big beam and hilltop location) the location was south of 30° latitude and that he had enough time to keep the band under constant surveillance. With that, we offer a salute to those dedicated chaps who keep beacon stations running in Rhodesia, Mexico, Hawaii, the Cook Islands, Australia, and Japan.

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ham radio



antenna and rotator preventive maintenance

Here's graphic proof that Ben Franklin's Advice about a pound of prevention is still true

Ted Woolner, WA1ABP, 30 Cedar Road, Shrewsbury, Massachusetts 01545

Most amateurs agree that the antenna system is the heart of any station. Much has been written on the emphasis that should be put on erecting the best possible radiating system to compete on the crowded bands. I'd like to add some footnotes to the antenna literature in the area of preventive maintenance.

The initial investment in an antenna and rotator is substantial. How many times have you heard fellows say on the air, "Sorry, my antenna's stuck on Europe. Big storm passed through, and I haven't had a chance to go up the tower and fix the rotator." With a little care and forethought, you can ensure reliable operation during the worst weather, even if you use the more inexpensive equipment.

The following notes explain how to avoid the disappointment and down time that can occur with stuck rotators and vibrating antenna elements during the winter months.

the antenna

It's pretty discouraging to come home, anticipating an evening of hamming, and find your pride and joy in bits and pieces all over the back yard. Beam antenna elements are usually made of aluminum alloy. Unless they're given the proper treatment, they just won't withstand the rigors of weather. Airframe manufacturers know this—aluminum

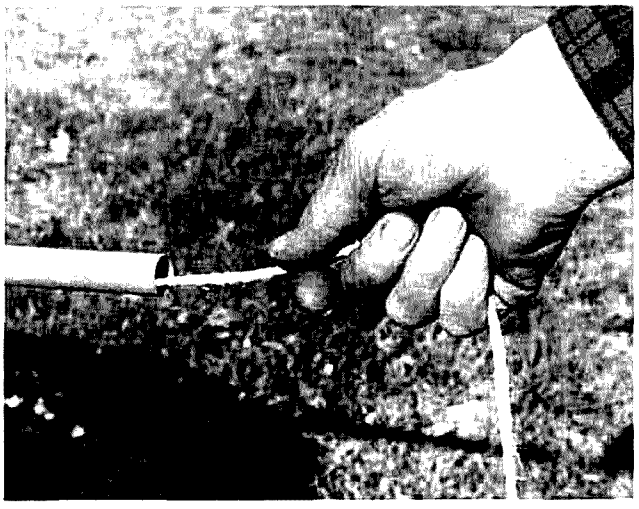
alloy is used extensively in this application. Where in the aircraft the aluminum is used determines the amount and type of treatment it's given before installation.

Most amateurs don't have the facilities for giving aluminum the correct treatment for withstanding environmental effects. The treatment is quite involved and includes chemical and electro-chemical techniques. It is quite expensive, also, if used for small quantities of metal.

One thing you **can** do, before hoisting your antenna to the top of your tower or mast, is give it several coats of zinc chromate paint. Put on as many coats as you can, and work it into every nook and cranny. For each coat of zinc chromate properly applied, you can count on at least six months of corrosion-free operation.

Now let's look at metal fatigue caused by vibration.

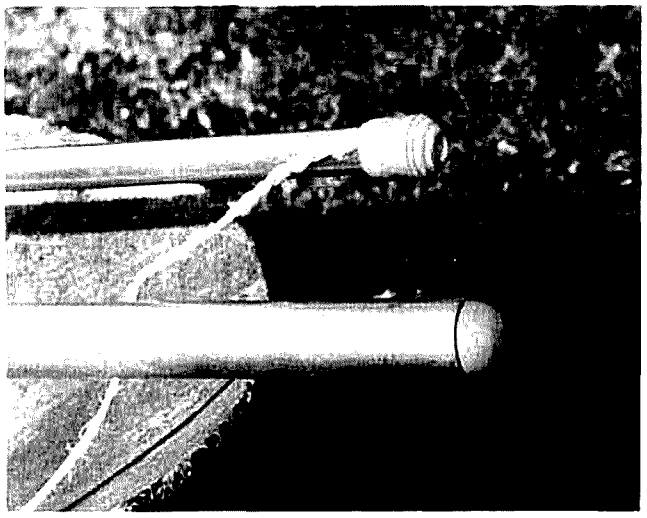
Tie a small weight on the end of the twine and drop it through the element.



taming the elements

No pun is intended in the heading of this part of the story. Every time your antenna sings in the wind, it's doing something other than playing a B-flat chord. Tiny cracks develop in the metal. They get larger, and as the element ices up, the added weight loads the metal. Vibration causes the metal to crystalize with resulting fatigue. Take a close look at your tubing the next time it falls to the ground after a storm. Those hairline cracks radiating from the break are a result of crystalization.

Hold the twine in place with a rubber foot, tennis ball or sponge rubber ball.



vibration proofing

All that's needed to dampen antenna element vibration is some sisal twine and something with which to plug the ends. Measure the length of all your elements plus a little extra, say a foot or so. Drop the twine through each element, and secure the twine with rubber caps over the element ends. A tennis ball is good. It may look a little strange, but who cares what the neighbors think? It's solid enough to make a good seal and will hold the twine securely. Make sure there's no moisture in the elements. The twine will dampen the vibration by changing the bending-mode frequency to some harmless low value.

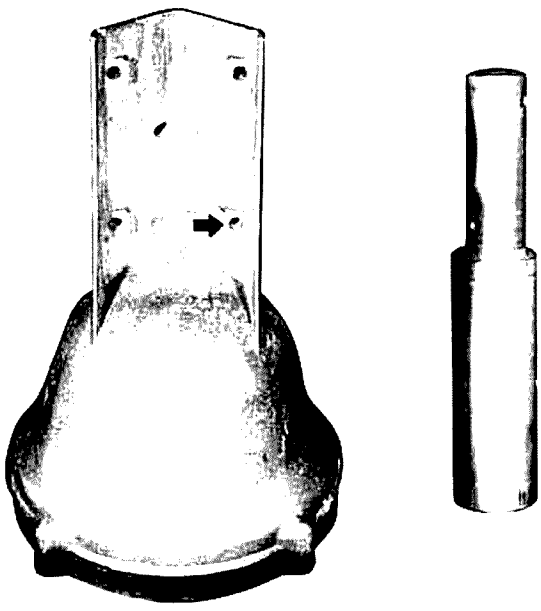
Another method is to pack the elements **loosely** with spun glass. Wear gloves when handling this stuff, though, because it can be pretty treacherous. It can penetrate your skin and cause all sorts of agony. Your boom can be sealed with a sponge rubber ball a little larger in diameter than the boom. Use the twine treatment here, also, and pull it up tightly. It will help keep your beam on a straight course and ease your jangled nerves

when your hear the wind wailing at 3 a.m. I'd like to emphasize that the spun glass **must** be packed into the tubing loosely. If it's packed too tightly you will zap the element.

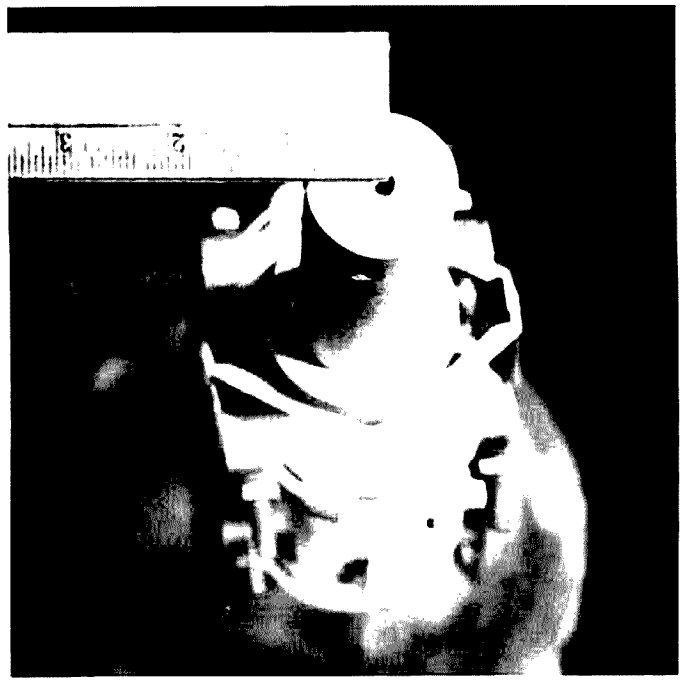
Another way of damping element vibrations is to fasten several rubber or plastic strips to the end of each element. Since the mechanical impedance of any rod-like element is high at the free end, an energy absorber at the end of an element will reduce the amplitude of vibration and minimize breakage. The energy absorber can be made from a five- or six-inch length of garden hose, split lengthwise every 90° and held at the end of the element with a hose clamp. These simple energy absorbers are most effective if they are clamped at the very end of each element, but to prevent changes in element electrical length and impedance, they should be placed so the free ends are about two inches in from the ends of the elements.

the rotator

Lots of stories have made the rounds about whether TV antenna rotators are strong enough to support amateur beams. The photos show my AR22 rotator, which was purchased, used, in 1963. At that time it had been in use for six months, and I've been using it constantly since. The reason it's down now is because I'm installing a new tower and two stacked five-element Cushcraft beams.



Rotator and reducer. The reducer is aluminum; large end is 2-1/16" in diameter and small end is 1½" diameter to fit inside mast. The holes running laterally were drilled through the mast to bolt the two together. Arrow points to one of four shoulders that may have to be filed off unless shims are used to center the reducer.



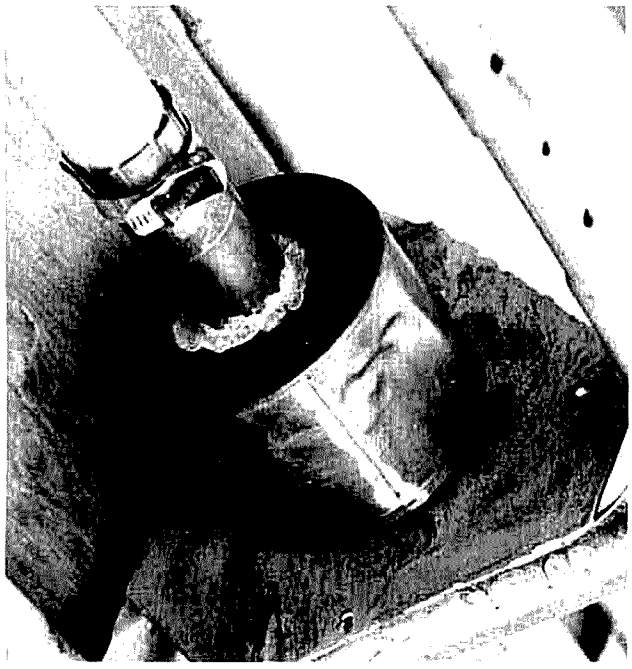
The top of the AR22 with reducer in place. The ruler was zeroed on the center. Right under the 1½-inch mark you can see the end of one U-bolt as it appeared with the rotor in the first direction; then it was rotated 180° and the U-bolts show again.

Given half a chance, I find that the AR22 rotator is very satisfactory. The secret is to eliminate the tremendous moment arm caused by a large mass of metal on a long length of mast above the rotator. Add to this the simple fact, sometimes overlooked, that any rotator must be perfectly centered with respect to its mast. Unless the correct diameter mast is bolted to the rotator, the eccentric rotation will load the rotator bearings, and you've got problems.

The AR22 manufacturer say that 1½-inch diameter pipe is correct. Maybe so, but I found this isn't true for my rotator. In my case, a diameter of two and one-sixteenth

inches is correct. I had a piece of aluminum stock turned down to this size, then had it stepped down above the rotator clamps to slip inside a 1½-inch mast.

It may be necessary to file down the edges



The hood or umbrella clamped on to the mast directly above the thrust bearing, which is out of sight and out of the weather; it's mounted on the plate welded to the tower cross-sections.

of the shoulders where the U-bolts come through to properly seat the aluminum reducer. I didn't have to, though.

The picture with the ruler has been purposely double exposed to show that the reducer is, in fact, centered in the rotator. I made the first exposure with the rotator pointed in one direction, then I turned it exactly 180 degrees and made the second exposure. Note that there's no visible deviation at the end of the reducer. If there were, two distinct centers would show, and two outer edges would also show.

the thrust bearing

A thrust bearing located somewhere between the top of the rotator and the top of the tower takes all the weight off the rotator. The only load on the rotator will then be turning torque, which is as it should be.

If the thrust bearing is exposed to the weather, it's going to get loaded with snow and ice. (California amateurs can skip this part. However, does anyone know what smog does to aluminum?) I made a copper umbrella to cover the thrust bearing, as shown in the photo. It slips onto the mast and is fastened with an ordinary automobile radiator hose clamp. The same idea can be used at the top of the tower where the mast enters the tubing.

Instead of a copper umbrella, you could wrap a piece of soft rubber around the mast and secure it with the same type of hose clamp. This will form a skirt to keep out the weather. Be sure the skirt flares at the bottom, else it might get loaded with ice and stick to the tower. It might help to lubricate the mast-skirt interface with low-temperature silicone to avoid this minor problem.

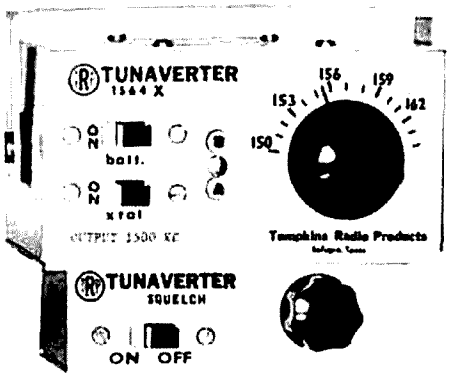
ham radio



"Yuh can't can't got thru a pileup like that . . . yuh gotta better chance out here on the edge."

new products

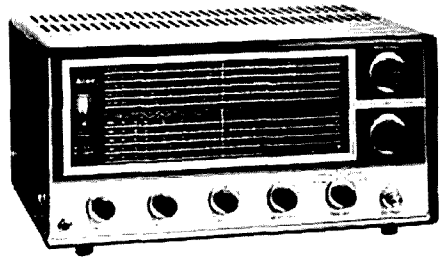
tunaverter squelch



Tompkins Radio Products has just announced a new squelch control for noise-free monitoring on hf and vhf Tunaverters. This makes it possible to have receiver squelch performance with an economical converter. The solid-state squelch circuit doesn't need any connections to the a-m broadcast radio **or** the converter other than the coaxial feedline.

This new unit features complete electronic control without relays and has a fully adjustable squelch setting. It is available in two models—the model ST that fastens to the bottom of a Tunaverter or the model SU which has its own mounting bracket. Furnished complete with connecting cable for $17.50 (model SU $1 extra). For more information, write to Herbert Salch and Company, Marketing Division of Tompkins Radio Products, Woodboro CTI, Texas 78393.

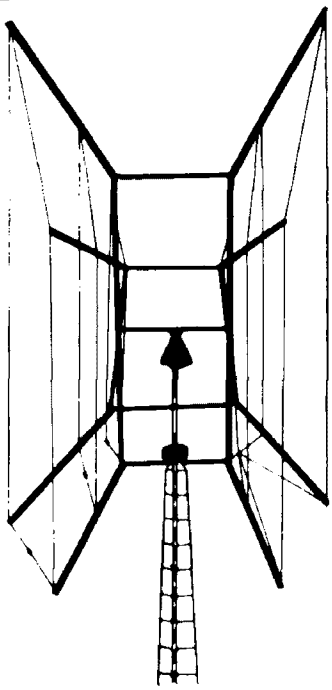
amateur and swl receiver



Allied Radio has recently announced a new solid-state receiver for amateurs and swl's, the A-2515. This new five-band receiver tunes all the amateur bands from 80 to 10 meters, international shortwave, aircraft, marine and the standard a-m broadcast band. The bands covered are 150-400 kHz, 550-1600 kHz, 1.6-4.8 MHz, 4.8-14.5 MHz and 10.5-30 MHz.

A total of 24 semiconductors are used in the circuit with two field-effect transistors in the front end for maximum sensitivity and low noise. Four mechanical filters provide sharp station separation. The noise limiter and agc reduce noise, blasting and fading. The built-in bfo and product detector give good reception of CW and ssb. The illuminated slide-rule dial has calibrated bandspread for the amateur bands; an S-meter is included on the front panel. Other features include a push-pull audio stage with a thermistor for low distortion, provisions for receiver muting and a headphone jack.

The receiver is equipped with dual power supplies, one for 117 Vac and the other for 12 Vdc, so the receiver may be operated from house current, autos, boats or camp sites. $99.95 from Allied Radio Corporation, 100 North Western Avenue, Chicago, Illinois 60680.



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new allied catalog

Allied's new 1969 catalog "Electronics for Everyone," is now available. This new book presents the latest in electronic components, vacuum tubes, semiconductors, wire and cable, test instruments, microphones and speakers, antennas and towers and amateur equipment as well as a full line of high-fidelity equipment and tape recorders, transistor radios, TV sets and video tape recorders. Specialty lines include a big selection of electronics kits, cameras and accessories, weather instruments and a wide selection of electrical accessories. The 1969 catalog, number 280, is available on request from Allied Radio Corporation, Post Office Box 4398, Chicago, Illinois 60680.

motorola semiconductor catalog

Motorola has just announced a new catalog of 175 semiconductors, books and accessories in the HEP line of devices for the amateur, experimenter and electronic hobbyist. The HEP line includes semiconductors at reasonable prices for almost any experimental or replacement use.

There are many new devices listed in this catalog: high-voltage germanium power transistors, complementary silicon npn/pnp power transistors, a unijunction transistor, a bilateral trigger diode, a sensitive low-power SCR, digital RTL and linear integrated circuits, an IC kit, new IC sockets and a TO-5 heat sink.

Other semiconductors in the HEP line include small-signal and power silicon and germanium transistors, field-effect transistors, silicon rectifiers and bridges, zener diodes, SCR's, power transistor mounting kits, heat sinks and books.

Copies of this handy new catalog, the Motorola HEP Catalog, MHA27-4, are available at your local Motorola HEP distributor or by writing to the Technical Information Center, Motorola Semiconductor Products, Inc., Post Office Box 20924, Phoenix, Arizona 85036.

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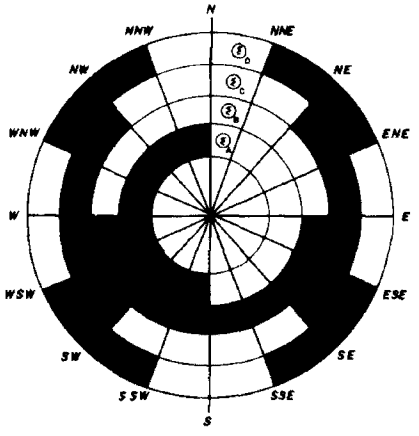
hammarlund ad

There's an error in one of the specs for the new Hammarlund HQ-200 receiver advertised on page 27 of the December issue—the a-m sensitivity should read 1 μV for 10 to 1 signal-to-noise ratio.

wind-direction indicator

The code wheels shown with the article on the digital wind-direction indicator in the September issue are in error. **Fig. 3,** shown as a Grey-coded wheel, is actually a binary wheel. **Fig. 2** is complete gibberish—a result of the photographic process used to make the illustrations for the magazine. A correct Grey-coded wheel is shown below.

fig. 1. Grey-coded direction wheel.



solid-state six-meter transmitter

In the schematic for the five-watt modulator on page 10 of the October issue, there should be a 270-ohm, 2-watt resistor in the line that runs from the center tap of the Stancor TA38 transformer (red lead) to the voltage supply. In addition, the line going to pin 9 of the integrated circuit should be connected to the two lines that cross it—one between pins 8 and 11 and the other from pin to the transformer.

calibrators and counters

In the decade counter shown in **fig. 3** on page 44 of the November issue, pin Q on flip-flop **C** should be connected to pin J of flip-flop **D.**

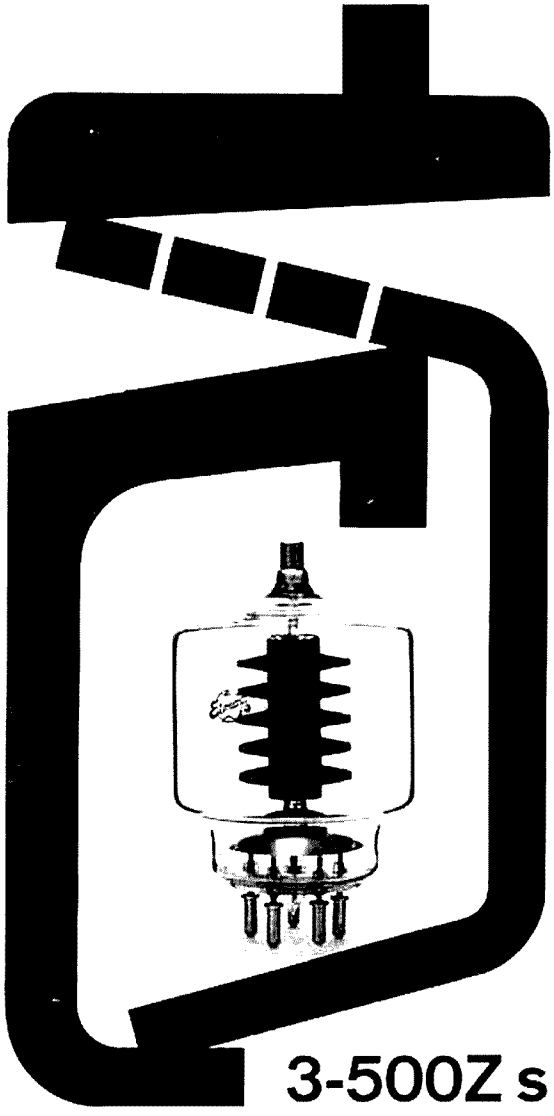
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editor
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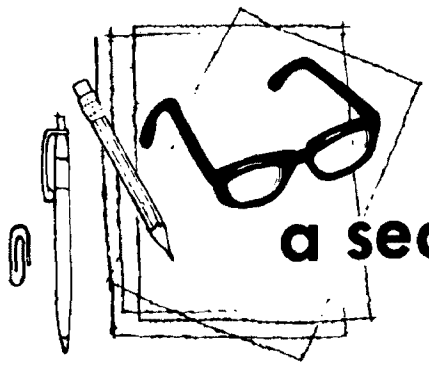
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a second look

by jim fisk

If you were asked to generate a generous amount of rf power you'd probably ask, "at what frequency?" For the high frequencies, it would be transistors or vacuum tubes; at vhf you could add varactors. For microwave you would tend to think in terms of large klystrons, magnetrons . . . or semiconductors that use drift time in their operation such as klystrons use *transit* time.

The first of the semiconductor devices to actually work was built by Ian Gunn at IBM in 1963. In the basic Gunn-effect device, a simple wafer of n-type gallium arsenide—without junctions—generates microwaves directly when a voltage is placed across it. When a constant voltage is applied to the material, the current through it fluctuates wildly at an extremely rapid rate, although somewhat randomly; however, if the semiconductor wafer is thin (about 0.005 inch or less) the current no longer fluctuates in a random fashion, but rises and falls in a cyclic way. Presto: microwave energy.

The fundamental frequency range of Gunn-effect oscillators is 5 to 35 GHz, although the Gunn device is not a narrow band gadget. On the contrary, the frequency is governed by the circuit. The active semiconductor material has a natural resonant frequency, so if it is imbedded in a completely resistive load, it will oscillate at its natural frequency. If the material is connected into a tuned circuit, it is possible to change the frequency and provide essentially a widely tunable oscillator.

What about power output? Not too bad; there are continuous-wave Gunn oscillators available which provide nearly a half a watt at 10,000 MHz. If you are willing to go to pulse, you can obtain about 50 watts at the same frequency with increasing amounts of power as you slide down the spectrum, going up to a kilowatt or so at 5000 MHz.

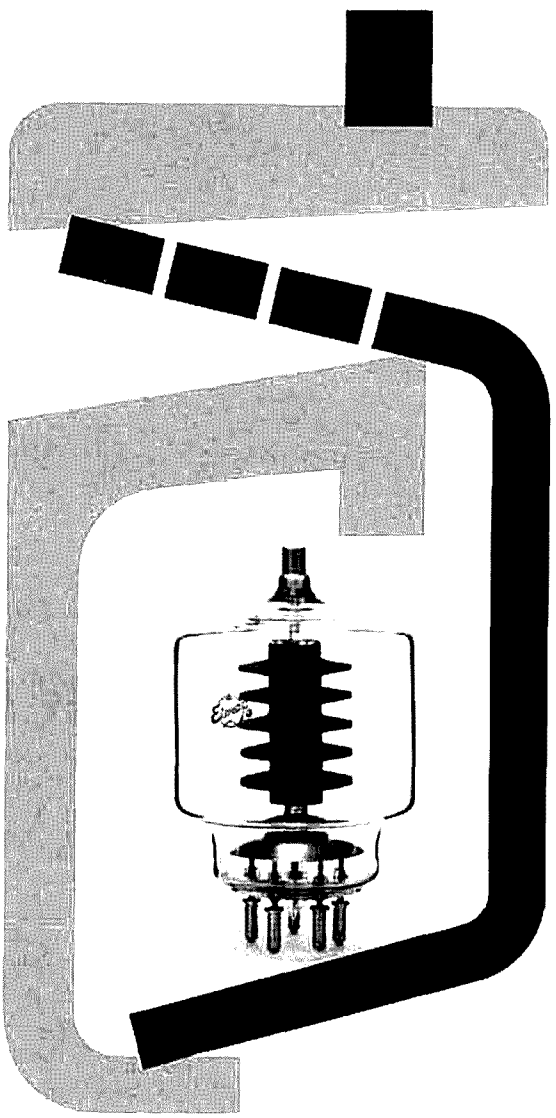
If you think the Gunn-effect device sounds interesting, consider one of the other microwave-generating diodes. The avalanche type or IMPATT diode (for IMPact Avalanche Transit Time) is just one; IMPATT diode oscillators operate because of a negative resistance that results from a combination of internal secondary emission and bunched current carriers (electrons and holes) that drift through the solid-state material and deliver rf power by causing an external circuit current 180° out of phase with the applied voltage. The main disadvantages of IMPATT devices are high noise and low efficiency—on the order of 5%. The CW output powers of IMPATT diodes are slightly higher than Gunn devices and pulse powers are about the same.

Although these diodes are bound to whet your appetite for more information, there is one device that is even more interesting; silicon avalanche diodes have been built that have several hundred watts output in the 400 to 1250 MHz range! Results are so good they do not agree with theory for these devices so they are *called*—most appropriately—**anomalous** diodes. As an example of a device of this type, at a frequency of 425 MHz, the output power was 435 watts with a 164-volt supply at 12 amps. This represents an efficiency of 22%, low by vacuum-tube standards, but the same diodes oscillated at 10,000 MHz (in the IMPATT mode) with their ordinary efficiencies of less than 5%.

So far I haven't mentioned cost—for good reason—it's high; so was the varactor when Sam Harris described the first practical parametric amplifier ten years ago, but that didn't stop interested amateurs from getting devices for their own experiments.

Jim Fisk, W1DTY
Editor

a grounded-grid two-kilowatt pep amplifier



This "full-house" linear uses a pair of 3-500Z's in a highly efficient circuit– W6SAI designs it for you and explains its operation

William I. Orr, W6SAI, Manager, Amateur Service Department
Eimac Division of Varian, San Carlos, California 94070

The popularity of the grounded-grid amplifier is well established. This interesting circuit combines high efficiency, good linearity, and minimum expense–three attributes not usually found together in one amplifier. A quick on-the-air check reveals that an amazingly large number of amateurs are using grounded-grid amplifiers and, judging from the robust signals, are having much success with them.

the grounded grid

To the uninitiated the grounded-grid or cathode-driven circuit is puzzling. A simplified sketch of this amplifier is shown in **fig. 1.** The driving signal is applied to the filament, or cathode, and the grid is at ground potential, thus acting as a shield between input and output. When the cathode is driven negative, the grid is positive with respect to the cathode, and plate current flows. The driving voltage is thus added to the plate voltage, and the resultant increase in power input to the stage appears as feedthrough power in the plate circuit. The plate circuit is conventional, being either parallel-tuned or the more popular pi or pi-L network.

At first glance it might appear that the grid receives no exciting signal. The driving voltage is developed **between the grid and cathode** of the tube, and it matters little which of these two elements is at ground potential. That is to say, the grounded-grid amplifier may be thought of as a grid-driven amplifier "standing on its head." This is not strictly correct, as other factors such as feedthrough power and grid-lead inductance enter the picture. However, if you understand the conventional grid-driven amplifier, you can understand the grounded-grid amplifier. The fact that the grid is at ground potential should cause you no distress.

If a well-shielded triode having a reasonably high amplification factor is used in the grounded-grid amplifier, neutralizing circuits and bulky bias supplies can be eliminated, reducing the amplifier stage to the basic essentials needed to make it function properly. The new Eimac 3-500Z, and its older companion the 3-400Z, are ideally suited to this class of service. They combine the characteristics mentioned above with modest cost, providing more watts output per dollar.

Of interest to the user is that the drive requirements of a grounded-grid, 2-kilowatt PEP amplifier using a pair of these tubes is of the order of the output power level of typical amateur exciters and transceivers; that is, in the 100-watt PEP region. Bulky and expensive power absorbers between amplifier and exciter aren't required. Best of all, a portion of the drive power appears as output in the grounded-grid amplifier. This almost repeals the ancient law that says "You can't get something for nothing." In this case, it is nearly true!

A pair of 3-500Z or 3-400Z power triodes will run comfortably at 2 kilowatts input

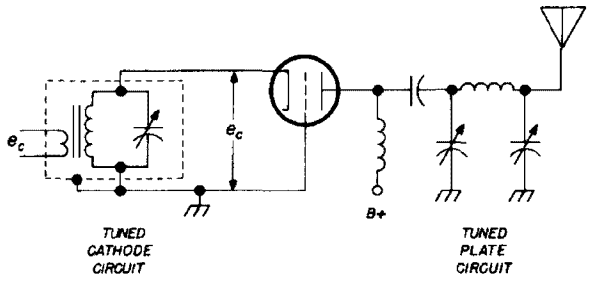


fig. 1. The basic grounded-grid amplifier. The driving voltage, e_c, is impressed between cathode and grid. The grounded-grid circuit isolates input from output and affords good shielding up to 100 MHz or so in tubes of the 3-500Z class.

PEP (1000 watts average power) under continuous ratings in CW, ssb or RTTY service without distress. "Fireman-quick" tune-up style, wherein the equipment must be tuned up in 30 seconds or less else the whole works goes up in a puff of smoke, is not necessary. This fire drill may have an element of excitement about it, but it can result in tube casualties when done in a

hurry, or under pressure, as in the closing moments of a DX contest. Adequate plate dissipation, on the other hand, is insurance against catastrophe from operator panic or error and provides a comfortable margin of safety that leads to equipment reliability and lower operator blood pressure.

two-kw pep amplifier

The Eimac 3-400Z and 3-500Z tubes are air-cooled power triodes designed for zero-bias, class-B rf amplifier service. They are particularly well-suited for ssb linear amplifier service. Intermodulation distortion products are 30 dB or better below one tone of a two-tone test signal for either tube type. A comparison of the 3-400Z and 3-500Z types is given in Eimac Amateur Service Bulletin number 35*.

A circuit for either the 3-400Z or 3-500Z is shown in **fig. 2.** The tubes are cathode driven with grids at both rf and dc ground potential. A tuned-cathode network achieves an optimum impedance match to the exciter and reduces intermodulation distortion to a minimum. The plate pi-network circuit is suitable for operation on amateur bands between 80 and 10 meters. Separate meters monitor grid and plate current. A VOX-operated circuit reduces plate dissipation during standby. Power output is better than 1200 watts PEP on all bands, and driving power is about 75 watts PEP. Plate potentials between 2500 and 3000 may be used, with reduced output for plate voltages as low as 2000.

the plate circuit

The amplifier plate circuit may be of several forms. The easy and more expensive assembly is to use a bandswitching pi-network inductor. The less-expensive but time-consuming procedure is to make your own bandswitching network from commercially available coil stock, copper tubing and a heavy-duty switch. Either approach will work well, and your choice will probably be made on an economic basis. Data for either type of circuit is given in the parts list, **table 2.**

One idea that hasn't been popularized is designing the amplifier for single-band oper-

fig. 2 (right). Schematic of the grounded-grid linear amplifier. RFC1 should be mountad reasonably in the clear so that the series-resonant frequency does not drop below 23 MHz. Capacitors marked C are disc ceramic, Centralab DD-103. Power-line bypass capacitors are Centralab DD16-103.

blower	**13 cubic feet per minute at 0.2-inch back pressure; use number-3 impeller at 3100 rpm (Ripley 8472, Dayton 1C180 or Redmond AK-2H-01AX)**
sockets	**Eimac SK-510 socket and SK-416 chimney for 3-400Z; Eimac SK-510 socket and SK0406 chimney for 3-500Z**

table 1. Cathode circuit components.

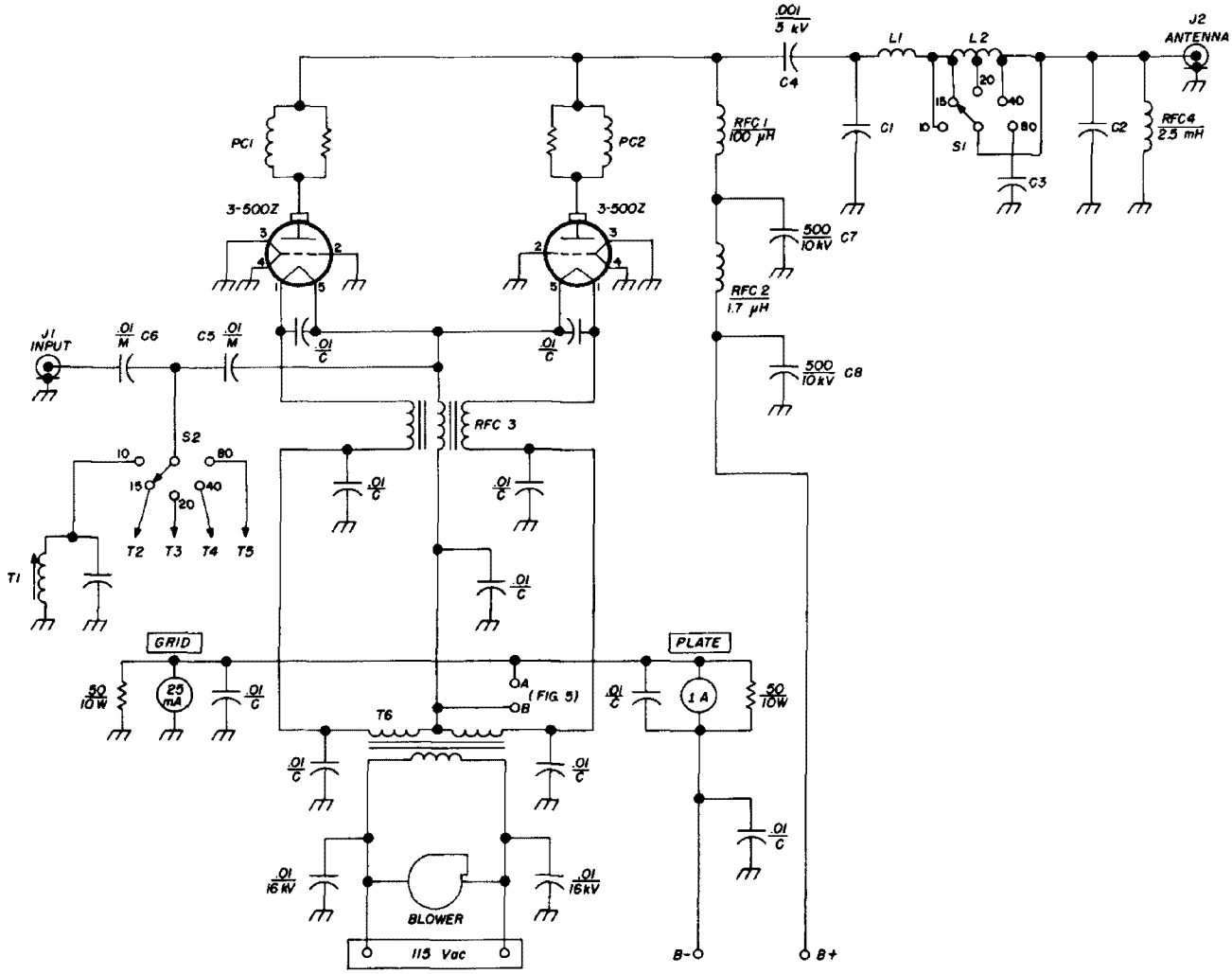
T1, 10 meters	**0.15 μH. 4 turns number-14 AWG enamelled wire on ½" form, ½" long (National XR-50 form); parallel capacitance: 200 pF, 1 kV silver mica; resonate to 28.7 MHz**
T2, 15 meters	**0.15 μH. Same as T1, resonate to 21.3 MHz with 470-pF, 1-kV silver-mica capacitor**
T3, 20 meters	**0.31 μH. 6 turns number-14 AWG enamelled wire on National XR-50 form; parallel capacitance: 470 pF, 1 kV silver mica; resonate to 14.3 MHz**
T4, 40 meters	**0.31 μH. Same as T3, resonate to 7.2 MHz with 940 pF (two 470 pF, 1 kV silver micas in parallel)**
T5, 80 meters	**1.3 μH. 13 turns number-18 AWG enamelled wire on National XR-50 form; parallel capacitance: 940 pF, as in T4**

table 2. Plate circuit components.

combination 1	**B&W 850A bandswitching inductor**
combination 2	**Air-Dux 195-2 pi-inductor**
combination 3	**homemade tapped inductor; 80 meters, 13.6 μH; 40 meters, 6.5 μH; 20 meters, 1.75 μH; 15 meters, 1.0 μH; 10 meters, 0.8 μH**

Combinations 2 and 3 should be tapped and trimmed to resonate as follows: 80 meters, 210 pF; 40 meters, 105 pF; 20 meters, 52 pF; 15 meters, 30 pF; 10 meters, 30 pF. Inductor should be parallel resonated and grid dipped with these capacitance values before placement in circuit. For combination 3, the 10-, 15- and 20-meter coil should be wound with 3/16-inch diameter copper tubing, and the 80- and 40-meter coil wound with number-12 AWG wire.

* Available from the author upon request.



C1	250 pF, 4.5 kV (Johnson 154-16)
C2	1500 pF (LaPointe Industries, Rockville, Connecticut, type J-1500-S-30)
C3	500 pF, 2500-volt mica (surplus)
C4	.001 µF, 5 kV (Centralab 858S-1000)
C5, C6	.01 µF, 500-V mica (two ElMenco CM-30 in parallel)
C7, C8	.001 µF, 6 to 10 kV (Centralab DD60-102)
L1, L2	see table 2
PC1, PC2	three 100-ohm, 2-watt carbon resistors in parallel (Ohmite "Little Devil"); three turns number 14 spacewound around one resistor
RFC1	100 µH, 1A (B&W 800) homemade versions: A. 112 turns number-26 AWG, spacewound (wire diameter) on 1" diameter, 6" long ceramic form (Centralab X-3022H insulator) series resonant at 24.5 MHz; grid dip with terminals shorted
	B. 136 *turns* number-20 AWG, 1" diameter, 4¼" long (B&W 3016 or Air-Dux 832T), slipped over varnished wood dowel for mounting, series resonant at 25 MHz; grid dip with terminals shorted
RFC2	1.7 µH (J. W. Miller RFC-144 or Ohmite Z-144) homemade: approximately 20 turns number-26 AWG enamelled wire wound on 100k, 2-watt resistor
RFC3	for series filament operation: 12½ bifilar turns number-12 AWG enamelled wire wound on ferrite core 3" long, ½" diameter (Indiana General CF-503, Newark electronics catalog number 59F-1521); notch core with file and snap to break
	for parallel operation: overwind third winding with number-18 AWG enamelled wire (see photo)
S1	single-pole, 12-position ceramic switch, 7 kV test (Radio Switch Corporation, Marlboro, New Jersey, type 86A)
S2	single-pole, six-position ceramic switch (Centralab PA-2001)
T1 - T5	see table 1
T6	series filament operation: 10 volts at 15 A (Thordarson 21F-146); parallel operation: 5 volts at 30 A (Chicago-Stancor P-6468)

ation. If your station operation (or the greater portion of it) is confined to a single hf band, a worthwhile reduction in cost may be achieved by constructing the amplifier for single-band operation. In this case, only one tuned input circuit is used, and the other transformers and switch S-2 may be omitted. The plate circuit may be simplified to a single tank coil, and the expensive and rather bulky plate switch may be omitted, along with the problem of placing the proper taps on the bandswitching coil. No hard and fast rules apply, and the choice is yours.

the input circuit

This is the feature that really distinguishes the grounded-grid amplifier from its conventional grounded cathode counterpart. A fixed-tuned cathode circuit is used, with each input circuit (T1 through T5) tuned to the center of each band (**table 1**). A grid dip meter is used for resonating the cathode circuits. Each circuit should be peaked before it's installed.

Because the cathode circuit replaces the grid circuit as the input system to the amplifier, the cathode circuit must be isolated from ground. This means that cathode voltage has to be applied through an rf choke.

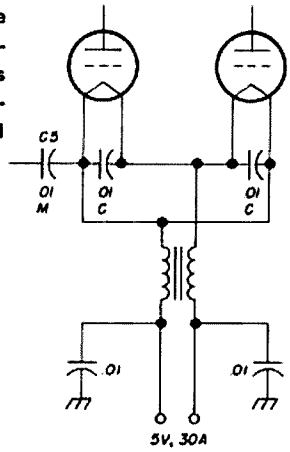


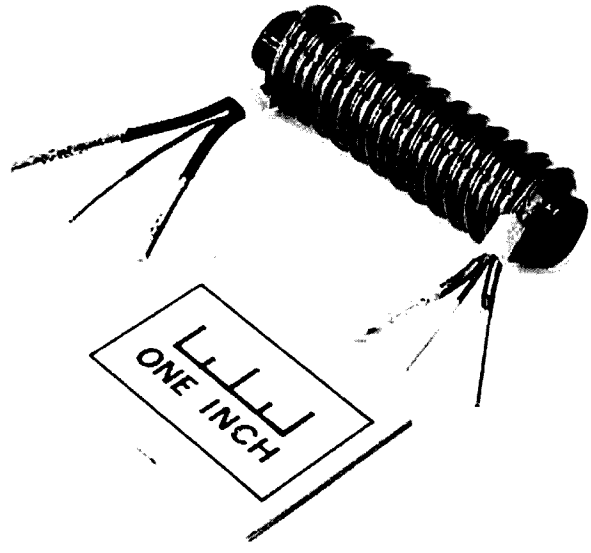
fig. 3. Filament choke for parallel filament connection. Core material is Q-1 type having a permeability of 125 at 1 MHz.

These chokes are available commercially, but are rather expensive. Why not make your own?

The filaments may be operated either in series, as in **fig. 2**, or in parallel (**fig. 3**). Series operation requires a current balancing return to the junction of the series filaments. This means a third winding is needed on the isolation choke. Here's how to make your choke.

the cathode choke

Use a ferrite core, but beware of the transistor-radio ferrites used for loopstick antennas. They **won't** work in this application and are inferior at higher frequencies. I used Q-1 ferrite material with a permeability ratio of 125 at 1 MHz. My core is one-half inch in diameter and three inches long. (The choke in the photo was wound on a two-inch-long core, which makes winding a bit tight; a slightly longer core makes the winding job easier.)



Filament choke for grounded-grid amplifier is wound on ferrite core, one-half inch in diameter and three inches long. The text explains how to build your own.

I used 12½ turns of number-12 AWG enamelled wire, bifilar wound, and held in place with epoxy cement. If you want to use series-connected filaments, wind another layer using number-18 AWG enamelled wire. Wind it over the primary. Epoxy cement will hold it in place. Tin the end of each winding and apply tape as shown in the photo. The assembled choke is extremely rugged when the epoxy dries, and you

will save at least $5.00 by making your own.

If the filaments are series-connected for 10-volt operation, the tertiary winding on the filament choke must make a connection between the center tap of the filament transformer and the junction between the series-connected tube filaments. If this connection is omitted, the filament voltages for the two tubes will not necessarily be balanced, as one tube may draw slightly more current than the other. Manufacturing tolerances

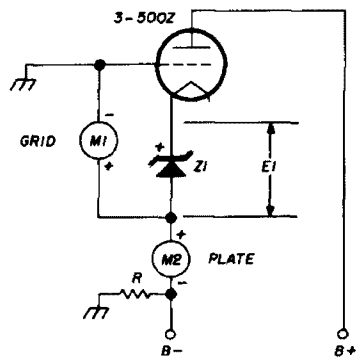


fig. 4. Bias and metering circuit for 3-500Z tubes. Meter M1 reads grid current between cathode and grid; meter 2 reads plate current between cathode and plate. The bias voltage, E1, is provided by the zener diode, Z1.

permit a current variation of about five percent between two tubes of this size. It's therefore possible for one tube to absorb the filament voltage at the expense of the other. Parallel filament operation eliminates this problem. However, 10-volt filament transformers are sometimes easier to obtain than 5-volt units.

the grid circuit

This circuit, by careful inspection of the schematic, is almost nonexistent. All three grid pins of each tube socket are grounded to the amplifier chassis. The proper Eimac air-system socket and chimney should be used, regardless of your choice of tubes. The less-expensive ceramic socket often used with tubes of this type applies too much lateral force to the base pins and glass header. The Eimac SK-410 socket has special, flexible connectors that allow the tube to move about. This equalizes any stress when plugging the tube into the socket. Also, this socket has small slots in the shell through which grounding straps may be passed to permit a short, low-inductance ground connection to grid pins 2, 3 and 4.

the metering circuit

You may wonder, "How is grid current measured when the grid is directly grounded?" The answer is that grid current may be measured in the grid return circuit to the tube filaments. Grid meter M1 handles this chore. The grid circuit electron flow, then, is from filament to grid within the tube, from grid to ground, from ground to the negative terminal of meter M1, through the meter, then from the positive terminal of meter M1 to the filament center tap.

The plate-current meter may be handled similarly. Instead of placing it in series with the B-plus lead, which may be highly dangerous, the meter is in the B-minus return lead to the power supply. No grid current flows through plate meter and no plate current flows through grid meter. This circuit is shown in **fig. 4.**

standby and vox control

The amplifier should be turned off to conserve power during listening periods and to reduce diode noise in the receiver. Under certain conditions, it's possible for large transmitting tubes in standby to generate a powerful hiss when they pass resting plate current. Normally, you don't hear this unless a T-R switch is used, but it can be a nuisance and is often noticed even when coaxial antenna changeover relays are used.

Common practice calls for the amplifier plate supply to run during standby, with the transmitter energized by the voice-controlled (VOX) system. A number of alternatives, then, present themselves to solve the diode noise problem and allow effective VOX control.

The schematic of **fig. 2** shows the filament circuit interrupted by two terminals (A-B). When these are open, the amplifier is in a cut-off mode, and no amplifier plate current flows. I don't recommend the amplifier be used under these conditions, but **fig. 5**

illustrates several acceptable circuits that may be used across terminals A-B to do the job.

Fig. 5A shows A-B shorted, in which the amplifier operates normally, with no VOX control and no provision for reducing noise. **Fig. 5B** shows the VOX relay contacts in parallel with a 10-k ohm, 10-W bias resistor. When the VOX relay is in standby, its contacts are open, and the resistor is in the filament-return circuit. A small plate-current flow through this resistor creates a voltage drop across the resistor. This applies a negative blocking bias to the amplifier grids. The standby plate current is thus reduced to near zero, and the noise is eliminated. When the VOX circuit is energized, the relay contacts short circuit the bias resistor. Amplifier operation then returns to normal.

Fig. 5C shows how a zener reduces amplifier standby plate current. At a plate potential of 3000 volts, for example, the 3-500Z and 3-400Z resting plate currents are 160 mA and 100 mA respectively. This corresponds to a plate dissipation of 480 watts and 300 watts. Unless VOX bias is used, this amount of dissipation per tube can warm the operating room and run up the power bill in short order. Addition of a zener provides a few volts of regulated bias, which reduces plate current to a moderate value. Standby plate dissipation is reduced accordingly.

Fig. 5D shows the use of both circuits. The VOX relay and resistor bias provide near zero standby plate current, and the zener provides low static resting plate current when the amplifier is activated.

If asked for my circuit preference, I'd vote for the circuit in **fig. 5B.** The addition of the bias, in my opinion, is "gilding the lily," unless the power supply is of marginal rating and every extra watt must be carefully conserved, or unless you want to reduce cabinet temperature during extended periods.

the air cooling circuit

Until some genius miniaturizes the watt, a means will be necessary to remove the heat generated in an amplifier of this size. When you realize that the filaments of the two tubes have a power capacity of 145 watts, and the total plate dissipation may reach **1000** watts during the tune-up process, you can appreciate that the amplifier cabinet temperature may reach astounding proportions unless the heat is conducted elsewhere. Large commercial transmitters that must deal with thousands, or hundreds of thousands, of watts of dissipation often forsake air cooling for more efficient water cooling or vapor-phase (steam) cooling. At the amateur power level, however, simple convection or forced-air cooling is still the most popular method of removing heat.

Either the 3-400Z or the 3-500Z requires cooling air supplied at a rate of 13 cubic feet per minute capable of overcoming a back pressure created by tube, socket and chimney expressed as a pressure drop of 0.13 inches of water. Blowers that will meet this specification are specified in the parts list. A word of warning: don't cheat on air supply. Heat is the enemy of vacuum tubes, and a war-surplus blower or cheap fan may result in decreased tube life due to filament

A. No bias and no vox; short circuit terminals A and B of fig. 2.

B. VOX-only circuit; relay contacts are part of external vox circuit.

C. Zener diode bias only; use 1N4551 zener diode for one or two 3-500Z's.

D. Vox circuit and zener diode bias.

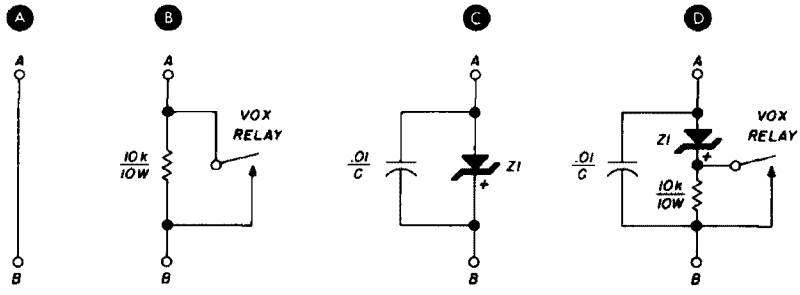


fig. 5. Bias and vox circuits for the grounded-grid amplifier.

or plate seals overheating. A few pennies saved in the purchase of the proper blower is false economy, indeed, and may lead to premature tube failure after a few hundred hours.

circuit stability

The grounded-grid amplifier is inherently a degenerative circuit, and the grid shields the input from the output circuit very effectively. The builder, however, is responsible for feedback occurring **around** the tube caused by coupling between input and output circuits. If sufficient coupling exists, it's possible for a fundamental frequency parasitic oscillation to occur. Proper amplifier shielding and correct isolation of power leads will eliminate this problem.

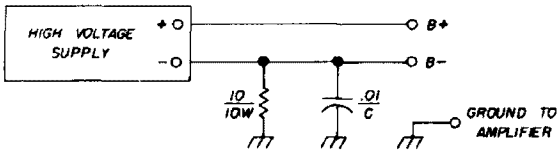


fig. 6. Power supply connections for proper "floating B-minus" circuit. Negative of supply is lifted above ground a few volts by the 10-ohm resistor.

The circuit shown in **fig. 2** has each power lead bypassed at the terminal board on the amplifier. Coaxial input and output rf connectors are used. If the cable run to the power supply is rather long, it's a good idea to run the leads in shielded cable. The B-plus lead can be a length of RG-8/U coaxial line, with the shield grounded at both ends.

The power supply B-minus circuit is above ground because plate meter M2 is in the return lead. A protective resistor should be in the power supply, as shown in **fig. 6,** to keep the supply from getting too far above ground.

The tuned-cathode transformers should be in a shielded enclosure to separate them from the rf currents and fields in the output circuit. There are several ways of doing this. One is to enclose the amplifier underchassis completely making it airtight. The blower is mounted on the rear wall of the enclosure, transforming it into a plenum chamber, with the air exhausted through the tube sockets. The cathode circuit, in turn, is automatically shielded by the enclosure.

On the other hand, if the underchassis area is not pressurized, and the air is conducted separately to each tube socket by hoses, the cathode circuit should be in a shielded enclosure of its own. The plate circuit components, usually mounted above chassis, should be in a shielded enclosure that affords proper ventilation. This can be done with a cover shield made of perforated aluminum sheet. With both grid and plate circuits properly shielded, minimum intrastage coupling will exist.

A good way to determine the amount and degree of intrastage coupling is to observe the action of the grid and plate meters after the amplifier has been tuned and loaded. If the stage is operating properly, maximum grid current and minimum loaded plate current will occur at the same setting of the plate tuning capacitor. If, on the other hand, grid current reaches maximum when the plate circuit is slightly detuned from resonance, it's an indication that unwanted feedback exists from plate to grid. A small amount of unwanted coupling probably will do no harm, but instability may often be found in such an instance, particularly on the 10- and 15-meter bands. If the degree of coupling is great enough, the amplifier will break into oscillation at the operating frequency. Judicious use of proper shielding can easily eliminate this problem.

parasitic oscillation

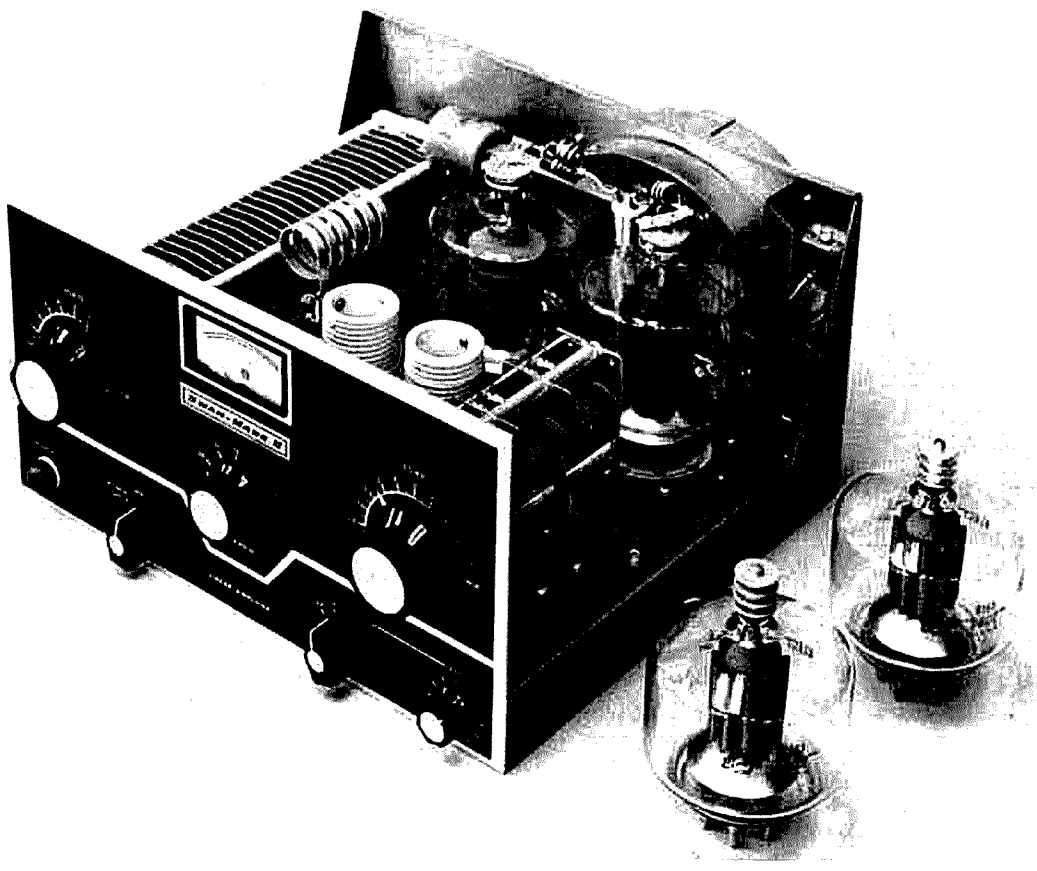
Most modern tubes perform well into the vhf region and demonstrate ample gain and efficiency, even though they may be designed primarily for high-frequency operation. The 3-400Z and 3-500Z are no exceptions. High-frequency circuit and construction techniques, however, usually create a condition wherein input and output circuits are not especially well shielded or otherwise isolated from each other at frequencies above 100 MHz or so.

This isn't surprising, as the amplifier is not designed to work in this portion of the spectrum. The upshot of this combination

of forces is that the tube is capable of amplifying (and oscillating) in the vhf region. Unless steps are taken to prevent it, this very oscillation will occur. Auxiliary neutralizing circuits may be added to suppress the tendency to oscillate, or, more simply, the parasitic circuit may be loaded to suppress the oscillation. This is most easily done by the use of plate circuit parasitic suppressors, a nearly universal adjunct to most high-frequency amplifiers. The parasitic choke described in the parts list should do the job for either tube type. One choke is placed in each plate lead near the plate connector.

The check for vhf parasitics is simple. The amplifier plate and filament voltages are applied, but excitation is not. Observe the resting plate current and grid-current, then tune the plate-tuning capacitor and loading capacitor at random. If parasitics are absent, the meter readings will remain unchanged. If, however, the amplifier exhibits a tendency toward parasitic instability, a grid-current reading will be indicated, and a rise in resting plate current will be noted.

When a new item of equipment is being tested for the first time in this fashion, it may be wise to drop the plate voltage in half with the aid of a variable voltage transformer in the power supply primary. If all looks well, the voltage may be raised a bit and the test repeated. The most drastic parasitic test of all is to boost plate voltage about 30 percent above normal while the test is being run. Higher-than-normal plate voltage raises the power gain of a tube and enhances the tendency towards parasitic oscillation, if such tendency is present.



New Swan Mark II linear amplifier, designed to use either Eimac 3-400Z (foreground) or 3-500Z (installed in amplifier). The Swan incorporates zener bias to reduce zero-signal plate current of the 3-500Z tubes; bias may be switched out for the 3-400Z's.

If a parasitic oscillation occurs, the plate circuit parasitic suppressors must be altered in unison until the amplifier is "cold." Addition of a turn or two to each suppressor coil is suggested as the first step. Too many turns on the coil will cause the suppressor resistor to overheat, particularly on the 10-meter band. A proper compromise between suppressor heating and parasitic suppression can be worked out without too much effort.

amplifier tuning and adjustment

Once the amplifier has been tested and is free of parasitic oscillation, it is ready for final adjustment and tuning. Using a grid-dip oscillator, the cathode circuits may be tuned to frequency before they are installed and will need no further tuning. A check of these should show each to be resonant at about the mid-frequency point of each amateur band.

Adjustment of the pi-network output circuit may also be approximated with the aid of the grid-dip oscillator. The output (loading) capacitor is set to the approximate value of capacitance for a particular band, the plate leads are attached to the tubes, and the plate tuning capacitor varied to provide resonance on a grid-dip oscillator coupled to the network coil. Settings of tuning and loading capacitors should be logged. A dummy antenna and power output device (an swr meter, for example) are attached to the amplifier, as is the exciter. Plate voltage is applied, and a **small** amount of drive (carrier) is introduced into the amplifier. The plate tuning capacitor is adjusted for maximum indication on the power output meter.

The idea is to achieve maximum power output with proper grid and plate currents, combined with a minimum of grid drive. This is how you do it:

As an example, assume the amplifier has a pair of 3-500Z tubes operating at 2500 volts. To achieve 2000 watts PEP input (usually assumed to equal 1000 watts under voice conditions) the amplifier **must be tuned and loaded at the 2000-watt level,** unless some rather sophisticated test equipment is at hand. Accordingly, 800 mA of plate current must be run to the two tubes, and the data sheet for the 3-500Z shows a grid current of 120 mA per tube, or a total of 240 mA for two tubes at this power level.

Here we go! Carrier is gradually inserted, and the amplifier is loaded toward the target plate current of 800 mA. Grid current is deliberately held on the low side as a safety measure. Plate loading is increased (pi-network output capacitance decreased), plate tuning is resonated and grid drive is slowly raised. At resonance we approach 700 mA plate current and 200 mA grid current. Tuning is "on the beam." Loading is juggled a bit to peak power output, and cathode drive is boosted a bit. The plate current is now 800 mA, grid current is 220 mA, and **power output drops if loading is either increased or decreased.** Power output also drops sharply when excitation is reduced. Under these conditions, the amplifier is loaded to 2000 watts dc input with the proper plate loading and with the currect ratio of cathode drive to plate loading.

Excitation is now removed, grid current drops to zero, and plate current returns to zero-signal value. Now, instead of a steady-carrier driving signal, a voice signal is impressed onto the amplifier. Voice gain is raised until voice peaks **occasionally** hit 400 mA. This indicates a "dc meter reading" of 1000 watts peak input (2500 volts at 400 mA). The PEP input, if viewed on an oscilloscope, would be in the neighborhood of 2000 watts or so. Grid current, under voice conditions, will peak about 110 mA.

Note that to establish a 2000-watt PEP input condition, the amplifier must be loaded and adjusted at the 2000-watt PEP level. This is extremely important; it must be capable of handling 2000-watt peak signals without distortion. It is wise to use an oscilloscope to monitor voice peaks and ensure that over-excitation or "flat-topping" of the signal does not occur.

The experienced operator will find that the combination of grid current and power output indications are important adjuncts to proper tuning. Once the proper tuning and loading technique is mastered, adjustment of a linear amplifier will be simple and uncomplicated.

ham radio

signal detection and communication in the presence of white noise

An informative treatment of the radio noise problem, with some answers based on detection theory

Peter Laakmann, WB6IOM, 8001 Airlane Avenue, Los Angeles, California 90045

Many years ago, when radio started, the detection of signals was receiver-sensitivity limited, since rf amplifying devices were not known. When these devices came about, it was found that receiving sensitivity was limited by natural and man-made interference rather than by the ever-present thermal noise of the antenna and its environment. This is still true for all but the higher amateur bands.

During the past several years new communications modes such as moonbounce, troposcatter and meteorscatter have become popular with pioneering radio amateurs. These modes have in common tremendous path losses ranging up to 280 dB, and all operate against a background of white (or thermal) noise. With receiver or antenna noise a limiting factor on one side and transmitter power and antenna size on the other, much brain power has been expended by amateur and professional alike to develop improved signal processing techniques and coding systems to achieve maximum performance.

Many claims have been made by amateurs for particular "circuits" that will pull signals out of the noise, "cancel" the noise, etc. Some of these claims are true, while others are oversimplifications and not ap-

plicable. The object of this article is to bring some of the findings of detection theory to bear on the problem and show what is not possible. I'll try to do this in simple language, with a minimum of math, but without yielding to the temptation of oversimplification.

the problem

Let's assume that we have reduced the incoming thermal noise to the lowest possible amount by low-noise antennas, feedlines and receiver front ends. What we've done is reduce the noise power spectral density in the frequency band of interest. This quantity is measured in watts per Hz of bandwidth. Let the problem now be: given this background of thermal noise density, how can we detect signals and communicate by modulating the signal in some way?

cw transmission

For ease of understanding we'll start with the simple proposition of detecting the presence of a transmitted rf carrier. Let's assume the carrier is transmitted at constant amplitude and constant frequency. At first glance it might appear that a bandpass filter of arbitrarily narrow bandwidth in the receiver would be the optimum choice for highest sensitivity, since the amount of noise in the filter bandwidth can be reduced to as low a value as necessary for detection of the carrier. Unfortunately this is not true.

Between transmitter and receiver exists a transmission path that always modifies the transmitted signal. Such phenomena as fading, scintillation, Doppler shifts, and multipath modify the signal in the direction of increased bandwidth. The best receiving system, then, is that which maximizes the signal-to-noise ratio (snr) for the received

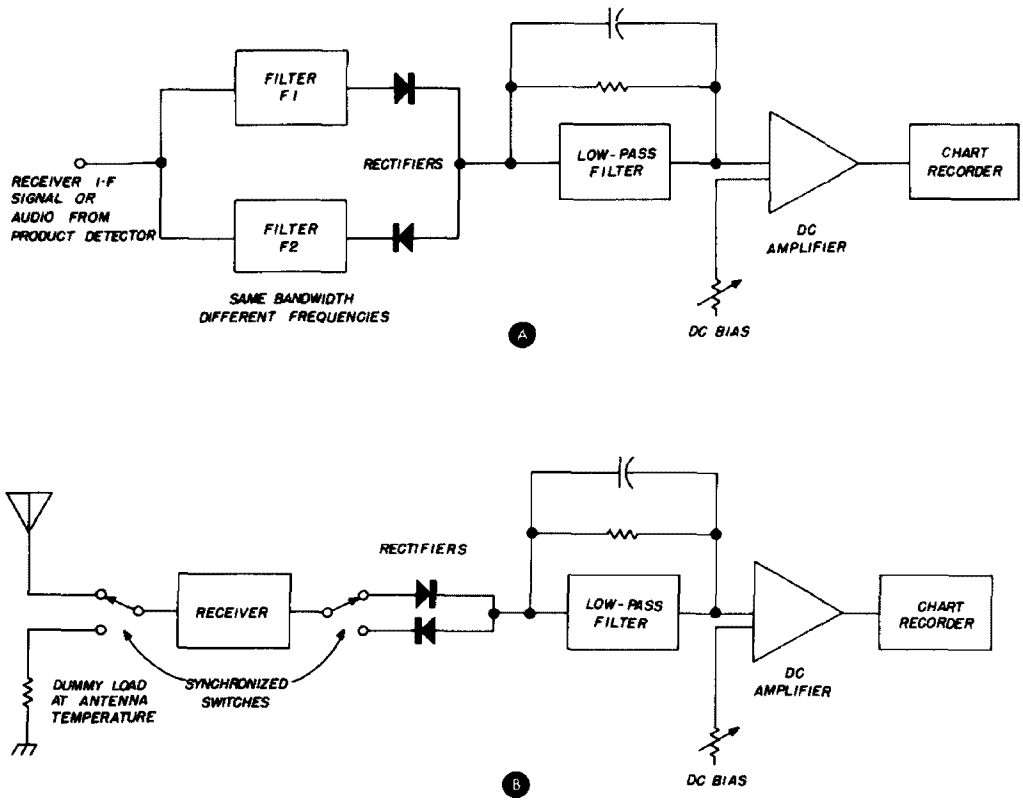


fig. 1. Weak signal cw receiving systems. The two-frequency method is shown in A; the Dicke radiometer system is shown in B.

signal; not the one being transmitted. A filter that maximizes the ratio of signal power to noise power is commonly called a matched filter. Such a filter for the above case would be one with a response approximately as wide as the signal spectrum; however, the exact frequency response of the filter is similar to the power spectrum curve of the signal being received. I won't go into the exact mathematics here, because we don't ordinarily know enough about a specific path at a given time to really shape the filter curve for optimum snr. Besides, the snr is not too sensitive to changes in the shape of the filter curve.

noise. Typically, the output of the filter is then displayed on a chart recorder or scope. The signal-to-noise ratio at the output of the low-pass filter can be computed by the following relationship:

$$\frac{S}{N_{out}} = \left(\frac{S}{N}\right)^2_{IF} \sqrt{\frac{B}{2\Delta f}}$$

$$(B >> \Delta f)$$

where

B = i-f bandwidth.

Δf = lowpass filter bandwidth.

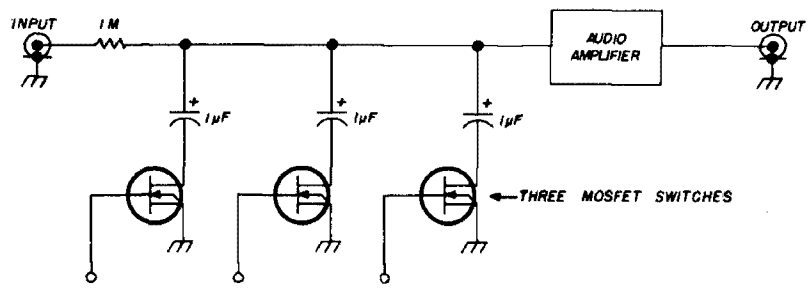


fig. 2. Electronic digital filter. A three-phase square wave at the frequency of interest is used to drive the three mosfet switches.

pre- and postdetection

Let's assume we've put into the receiver i-f this approximately matched filter, yet the signal-to-noise ratio is not high enough to detect the cw carrier. Can anything be done to improve snr using additional processing? Yes, but the price paid for improved detection capability is a drastic increase in *time-to-detection*. The additional processing techniques are called postdetection integration.

Basically, they all involve reading the long-term value of the noise and thereby detecting small increases of power caused by the presence of extremely low-level signals. They are nonlinear systems, and the output signal-to-noise ratio is not proportional to the input signal amplitude.

The implementation of postdetection systems is always to rectify the i-f, rf, or audio (in a heterodyne receiver) and smooth the resulting dc in a low-pass filter. It's intuitively apparent that the longer the time constant of the filter, the easier it is to detect a small change in average signal plus

Note that the i-f signal-to-noise ratio appears as a square term in the equation, so that even very large values of integration times produce only a small reduction in permissible i-f snr. The time-to-detection can go clear out of sight before there is a significant output signal-to-noise ratio or improvement in sensitivity.

While the equation is exact, it has to be applied to a system with care. If the signal amplitude is not constant you can't put in a fixed number for the i-f snr. To be exact you must integrate over a period of time. The low-pass filter bandwidth in a system involving a recorder and a human observer is really the combination of the electrical filter and the brain of the observer looking at the chart.

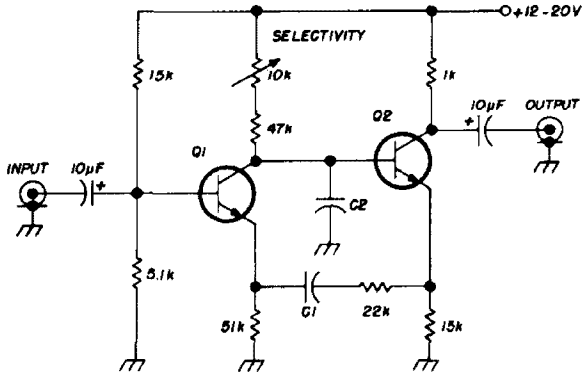
In a practical system the observer may be contributing most of the filtering.* I've found with systems of this type that there is little point in using integration times over 2 seconds on a cw carrier. With rapidly fad-

* And, depending on the observer, some of the noise.

ing moon echoes I prefer an integration time of 0.5 second, doing mental filtering on the chart output.

As a point of reference I estimate that the signal path contributes about 10 to 20 Hz of additional bandwidth on amplitude modulation. As an example, a cw system designed for highest sensitivity on moon echoes would use a predetection filter with bandwidth appropriate to the band used (20 Hz on 1296), followed by a postdetection integrator of between 0.1 and 0.5 second, followed by a recording device and the operator's mental filter. This is the optimum technique, and no fancy gadgets or trick circuits can do better than that.

Regenerative audio filter. Transistors Q1 and Q2 are high-gain silicon npn units. C2—about 0.05 μF—is one-half the value of C1 and depends on the desired audio frequency.



practical considerations

In order to keep small drifts in receiver gain from affecting the recording, most people use two different audio filters and two rectifiers in a bucking arrangement working into an integrator, so that the output is normally zero in either filter with no carrier. Since the noise output from the two filters adds, while the dc component cancels, the snr of this system is 3 dB worse than that with the single i-f filter. This is well worthwhile, though, to obtain system stability.

If frequency-shift keying is used, performance is identical to a keyed carrier in a single filter. A less-popular technique is to have a reference noise source at the receiver input and synchronously switch it and a pair of rectifiers at the integrator input to cancel receiver gain drifts. This is the technique used by radio astronomers. For cw use it's more difficult to implement and no better in sensitivity. (See **fig. 1**).

A disadvantage of the Dicke system is that it can't discriminate against radar interference. The two-frequency technique does discriminate, because the radar pulse adds to both channels equally if they have the same bandwidth.

predetection filter schemes

The two-meter moonbounce enthusiast might wonder how to implement a two-to-five-Hz bandwidth filter for the predetection (i-f) circuit. There are several possible methods:

1. Use a circuit similar to a Q multiplier, but at audio frequencies. This requires well-regulated supplies and a temperature-stable transistor circuit. Any transistor oscillator circuit will do if adjusted for class-A operation and if the feedback is adjusted to a point just before oscillation.

2. Use a passive RC circuit; with good components and regulated supplies, this circuit is capable of about 1-Hz bandwidth. Its operation is similar to that of the Q multiplier concept in method number **1.**

3. Use a digital filter (shown in **fig. 2**). This is the ac equivalent of a synchronous detector. It's extremely stable and can generate an arbitrary narrow response. I've built some of these with 0.1-Hz bandwidth at 1 kHz. Their resonant frequency is that of the ac source driving them. Some care is required in layout to avoid coupling the oscillator output into the amplifier input. Their Q depends on the basic cutoff frequency; the lower the cutoff frequency, the higher the Q. This network can't oscillate by itself.

4. Use a synchronous detector followed by a low-pass filter. This is equivalent to another heterodyne process, translating the high audio frequencies to near dc, where they can be filtered without high-Q devices. This is a

very useful device if phase coherence exists between the drive source and the carrier. It is more difficult to use than method number **3.**

Other considerations in choosing your predetection bandwidth are transmitter and receiver stability and the search problem with a too-narrow system. At 1296 MHz, 50-Hz bandwidth is probably as low as you can go and be sure the carrier is centered in the passband. At two meters a system of a few Hz bandwidth should be practical.

Fig. 3 shows a system that I am using on 1296 MHz. On moon echoes it is about 3 to 6 dB more sensitive than the human ear, with a 0.5-second integration time. **Fig. 4** shows typical moon echoes with a 0.5-second integration time.

pulsed systems

Let's now consider pulsed systems. Can anything be gained by transmitting a high ratio of peak-to-average power? The answer is a qualified "yes" for the relatively low power in amateur transmissions.

In general, pulsed and cw systems can transmit the same information rate if both are limited by white noise, and both use the same average power, provided each system is not limited by path a-m or fm. Both pulsed and cw radars perform identically with respect to maximum range and time-to-detection if both use the same average power.

However, for very high average power, transmission lines and antennas can't support the energy without breakdown. It's much easier to handle high average power if the peak-to-average power ratio is small. For these reasons new, long-range, high-performance radars use the cw mode. Range and range rate data are obtained by pulse compression at the receiver.

These radars are sometimes called "chirp radars." They transmit a "long" pulse of linear frequency. At the receiver the long pulse goes into a special filter that has a delay which also changes linearly with frequency to generate a short, high-peak power pulse. This overcomes the peak power limitations of the antenna and feedlines. These filters are very critical and are not yet within the realm of amateur work.

To show the equivalence of the cw and pulsed modes, let me go through an example. Assume two transmitters of equal average power output, no path modulation, and both transmitters operating for a 1-second interval. Allowable time-to-detection is one second.

Since we have an allowable "waiting time" of one second, corresponding to an information rate of one bit per second, the best pulse repetition rate is one pulse per second, because signal power is maximized in each pulse.

The optimum cw bandwidth for a 1-second "long" pulse is 1 Hz. (This is again the bandwidth of a matched filter.) It can be shown[1] that the shortest pulse length an amplifier can produce as useful output is equal to 1/B seconds, where B is the bandwidth in Hz. Conversely, the bandwidth for a pulse of any length, L, that will maximize the snr is equal to 1/L.

For the pulsed transmitter, the bandwidth required for maximum i-f snr varies proportionally with the peak power and inversely with the pulse length. As the noise power increases in the i-f bandwidth, the i-f snr is constant for any pulse length between zero and one second. On the surface one might think that pulse transmission can offer a tremendous increase in snr for a given average power level. Short-duration, high-power pulses could be used in conjunction with a gate to remove the noise during quiescent periods, resulting in a high snr during gate-on periods.

In the real world, with path distortion due to a-m or fm, it's a different matter. If the receiver bandwidth is chosen on the basis of the path rather than the desired information rate (as in the above example) there is much to be gained.

For the 1296-MHz system, we arrived at a bandwidth of 50 Hz for optimum detection. On the basis of information rate, however, we could live with a system transmitting one bit per second. For these conditions, then, a system transmitting one pulse per second with a pulse width of 20 milliseconds would be about optimum. This system would have a 17-dB improvement over the cw threshold.

It may be easy to generate such long pulses, but there is also a question of legality for that amateur frequency. However, this does show what is possible. The mode would be legal on 2300 MHz and higher bands. The bandwidth of this system is no larger than that required for high-speed cw.

further enhancement of snr

The comments made earlier about post-detection systems still apply to pulse transmission. If the i-f snr is not high enough with the 17-dB improvement due to pulse transmission, the pulse can be integrated to enhance the output signal-to-noise ratio. It's desirable to gate out the noise between pulses, as those intervals don't contribute to the signal. The calculation of output snr is:

$$S/N_{out} = (S/N_{1\text{-}f})^2 \sqrt{N}$$

where N = number of pulses integrated.

There is one other consideration with pulse transmission. The moon is of considerable size, and if a short pulse is sent up, the pulse will arrive at different points on the moon at different times. The reflected pulse is therefore stretched out in time. It turns out that the reflected pulse width is about 100 μs, which corresponds to a reflection from the near parts of the lunar disc only. Also, there's a very much longer tail that corresponds to the full radius of the moon; however, the tail doesn't contain much power.

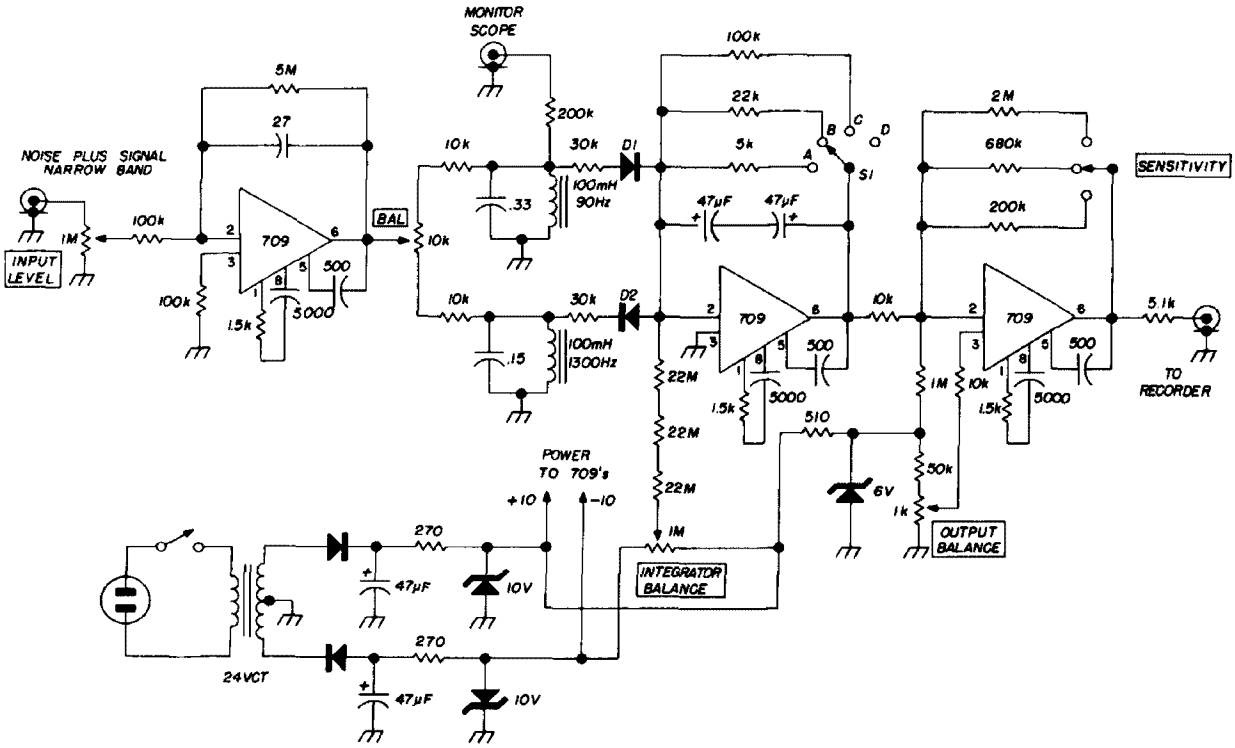


fig. 3. Versatile postdetection system used at WB6IOM. The positions of S1 are A, low sensitivity, fast response (0.1 second); B, high sensitivity, slow response (0.5 second); C, maximum sensitivity, slow response (2 seconds). D is the echo accumulator position.

bandwidth reduction on moonbounce

It would seem, then, that a bandpass filter of about 10 kHz bandwidth at the receiver should result in optimum snr, but this isn't necessarily so. If the transmitted pulse has a width of 1 μs, corresponding to a bandwidth of 1 MHz, the returned pulse may be ex-

panded in time without a corresponding reduction in bandwidth due to the incoherent nature of the moon reflection.

It's not clear, from any of the papers I've read on the subject, as to how much of a bandwidth reduction does occur, if any. If no significant reduction in bandwidth occurs, pulses less than 100 μs wide aren't desirable since they actually force the receiver bandwidth up without the corresponding benefit of a short, high-peak power pulse to separate signal and noise.

If a considerable compression of bandwidth occurs due to the physical size of the moon, systems using short pulse widths may be as efficient as systems using longer pulses. There are a variety of planar triodes around that can generate power at good efficiency when pulsed as 2300-MHz amplifiers, so this appears to be a useful approach to low-cost power generation. It would even be possible to send short pulses spaced less than 100 μs and get essentially cw returns. However, in the absence of hard facts relating to the phase coherence of the moon at these frequencies, this must await expert opinion.

voice transmission

The optimum transmission mode for voice signals against a background of white noise is still surrounded by a lot of confusing statements, even in the professional literature. However, properly defined and stated, it is a closed subject. For amateur purposes we are interested in a system that requires the lowest amount of signal power at the receiver for good voice intelligibility; not high output signal-to-noise ratio. Some possible transmission modes to consider are single sideband suppressed carrier, a-m, narrowband fm, wideband fm, pulse code modulation, pulse amplitude modulation, pulse position modulation and others.[2]

If voice transmission systems are compared on the basis of peak and average power, ssb is the obvious choice. However, a-m is superior to wideband fm and equal to narrowband fm, on the basis of the same carrier power.

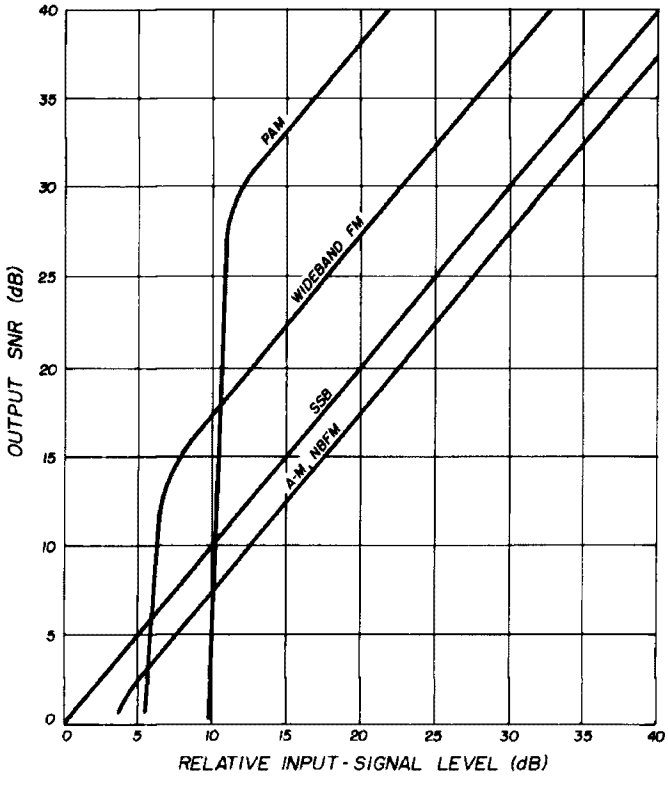
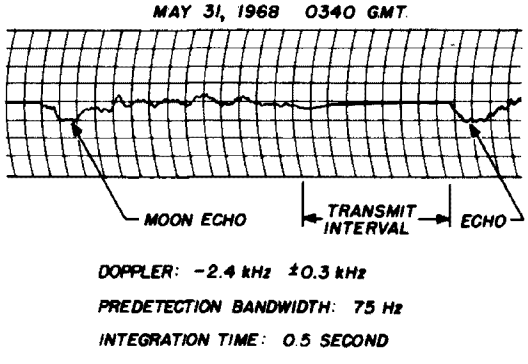


fig. 5. Signal-to-noise ratio characteristics of various modes of communication.

The advantage of fm, PCM, PAM, etc., is that these systems have noise improvement properties. In all these systems, the output snr may be higher than that of the carrier. The price paid for this is a higher signal threshold below which there is no intelligibility. **Fig. 5** compares the different systems qualitatively on the basis of equal input power to the receiver.

While the noise improvement property of

fig. 4. Chart recording of moon echoes received by WB6IOM on 31 May 1968.



fm, etc., is important for data transmission and high fidelity entertainment, it doesn't help the voice intelligibility very much. On the graph of **fig. 5** the voice intelligibility is already near 100 percent before the fm or pulse curves result in a larger snr than ssb.

in summary

Whether you listen to weak cw signals by ear or use a postdetection system and detect the signal by eye, it's extremely important to keep the system linear. The human brain is a very complex signal processor that can adapt itself optimally to a variety of signals, *but only if you don't* make decisions for it with relatively unsophisticated devices like level thresholds, keyed oscillators, or clipping circuits. Avoid all chances of clipping in the system, and particularly avoid level thresholds as in some popular "black box" devices.

The philosophy here is simply this: if you put a circuit in your black box that makes decisions about the presence or absence of signals, then you have presented your brain with a problem about which it has no choice other than to reject or accept the black box decision. This is probably not too bad if the black box contains several sophisticated computers, but it's not in the realm of anything as simple as, say, a couple of transistors.

By the way, it's quite difficult to build any system that's better than the human ear listening to an audio note in white noise. The ear acts like a tracking filter with a tracking range of about 3 kHz and an instantaneous bandwidth of 25 to 50 Hz. From about 200 Hz to 1 kHz the ear's sensitivity is constant and will detect signals with unity signal-to-noise ratio in a 25-50 Hz bandwidth (**fig. 6**). For signal detection of a carrier in white noise, about the best one can do is use a *maximally flat (less than 1-dB ripple) 100-Hz* to 1- or 2-kHz bandpass filter, with earphones or speaker that are also flat with frequency. Small resonances or peaks are quite distracting, as are some of the shenanigans now going on in the sub-bands reserved for extra class amateurs who are trying to work DX on these frequencies.

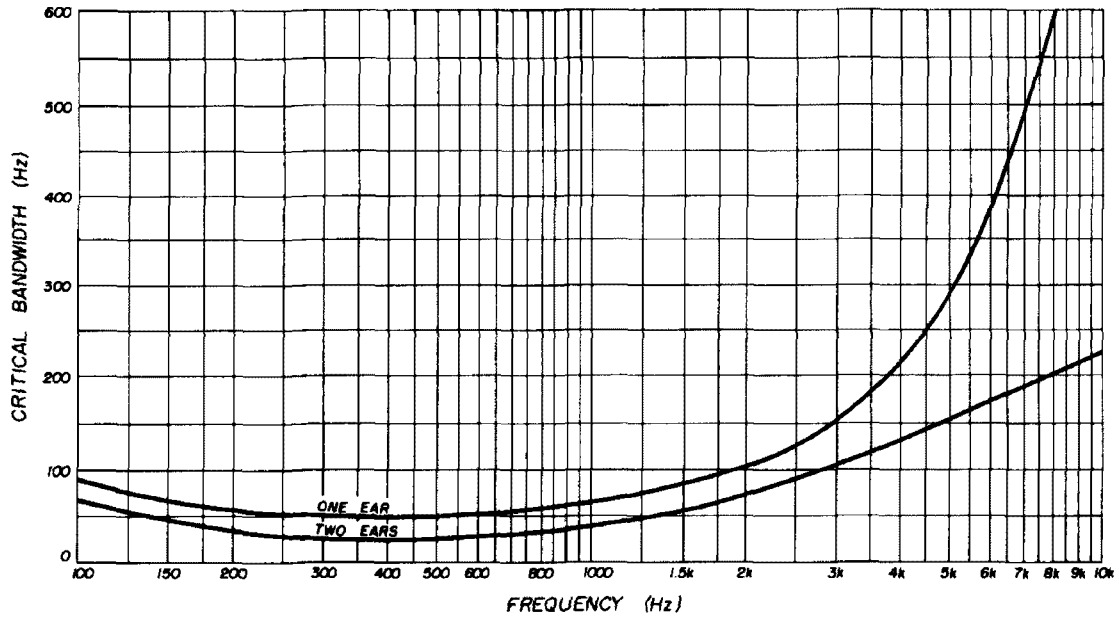


fig. 6. Critical bandwidth of the human ear as a function of frequency.

It's a good idea to check your ear for its sensitivity peak, because this peak can vary from about 200 to 600 Hz, depending on your age. Narrowband audio filters can be very tiring on the ear, because signal and noise begin to sound alike after an extended

period of listening (again depending on age of the listener). These filters are really no more sensitive than flat, wideband filters.

If you are an experienced weak-signal operator, you'll also have noticed that the brain can make a better decision about the presence of a signal if the signal's frequency is changing very slowly. (Any DX'er will certainly recognize this as a subtle phase shift between signal and noise; it requires extreme concentration, however.)

Audible detection of a signal, and copying it in the form of code, are two different things. The signal has to be at least 3 dB above threshold and peaked at your particular aural resonance point before you can decipher intelligence. This is because the *minimum* change in signal strength detectable by the human ear is 3 dB. The advantage of the postdetection scheme is that you can get positive copy, whereas the ear can just barely detect the presence of a carrier.

Another point to remember, when using ear filtering, is that the bandwidth of the ear is very dependent on volume; maximum sensitivity is usually at a relatively low volume when copying ssb and cw signals.

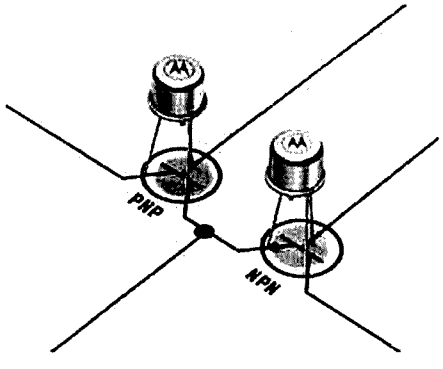
Carrier and cw detection by means of the ear are almost as sensitive as the best postdetection system known. They have the advantage of being much less sensitive to frequency instability and tuning errors. However, full sensitivity can only be realized if the system is treated with about the same care as a sophisticated high-fidelity receiving system. A good system should have separate volume adjustment for each ear, some indication of noise level for each ear, and at least two filters of 400 Hz and 2 kHz bandwidth. Optional would be adjustable frequency shaping to compensate for the different response of the two ears.

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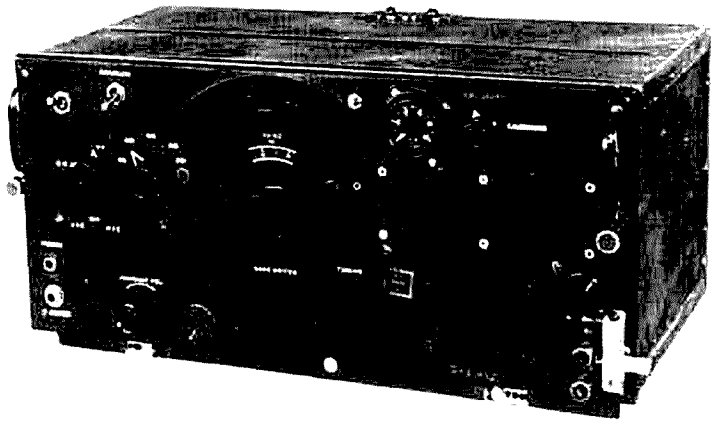
ham radio

large signal vhf/uhf transistor



Motorola has just made the first silicon PNP large signal vhf/uhf power transistor available to the electronics industry. Designated the 2N5160, it is the forerunner of several PNP transistors intended to compliment existing NPN devices. The 2N5160 was designed specifically as the PNP compliment of the NPN 2N3866 for use in complimentary circuit configurations such as vhf and uhf amplifiers.

This new transistor is very helpful where a positive ground must be maintained and eliminates many problems with bypass capacitors. Other uses for these power transistors are in unsaturated switch applications at higher current levels than are presently available with PNP switches. This transistor features a minimum power gain of 8 dB with no emitter tuning and 1-watt minimum power output at 400 MHz. It is suitable for use in class A, B, or C output, driver or oscillator applications at frequencies up to 800 MHz. The 2N5160 uses the multiple emitter overlay geometry and is available in the TO-39 package. For further information and complete specifications, write to Motorola Semiconductor Products, Inc., P. O. Box 13408, Phoenix, Arizona 85002.



converting a vacuum-tube receiver to solid state

You can reduce drift in older receivers by an order of magnitude— a modification scheme for the BC348 using MOSFET's and JFET's

Hank Cross, W1OOP, 11 Birds Hill Avenue, Needham, Massachusetts 02192

The **BC-348 receiver** is well known for its excellent selectivity and mechanical stability. At the same time, this old iron box can be an instrument of torture for the user. Some of the earlier BC-348's were bedeviled with the "scratchy dial" syndrome. This is a phenomenon of unknown origin and is virtually impossible to cure. It appears when the receiver is tuned and sounds like a thousand demons rubbing a tin washboard. If you're unfortunate enough to have a BC-348 with this ailment, and the receiver has a drift problem, it can be a real chore indeed to keep a signal in the i-f passband.

In resuming an interrupted conversion of the BC-348 begun a dozen years ago, I decided the only way to reduce frequency drift was to reduce internal temperature rise. There are several ways to do this. Reducing B plus takes off seven watts; a transistorized audio circuit eliminates another nine. To reduce drift by ten times, however, the power consumption must be cut by 90 percent. Answer: **all** transistors.

If the conversion includes a class-B audio amplifier, drain under normal conditions would be well under ten watts, and two watts of this would be in the pilot lights. Fine. Now the question is, how to do this without sacrificing any of the features this receiver is noted for?

choosing replacements

The BC-348, -312, and -342 receivers use rf coils with low-impedance plate windings (or taps) and high-impedance grid windings. Bipolar transistors might be used by turning things around (interchanging grid and plate), or junction FET triodes might be used in a cascode circuit, as in the H. H. Scott fm tuners—two JFET's replace one 6K7.

However, the RCA dual-gate MOSFET's, recently put on the market at a reasonable price, have the correct input and output impedances for this use and afford a gain control with lower distortion products. It turns out that the 3N140 MOSFET, at rated current, has about five times the transconductance of the 6K7. The relatively low output impedance of the 3N140 (**table 1**) is not objectionable when used with low-impedance inputs.

Because of the interstage coupling, stage gain in the BC-348 receivers increases from the low to the high end of each band. In some receiver models a gain control potentiometer was ganged to the tuning capacitor. I felt that this was an added complication not needed for amateur work, so I cut the wires going to the pot and those from the pot to the bandswitch inside the coil box.

Tubes I replaced with FET's were the 6K7 rf amplifiers, a 6J7 mixer, and 6C5 local oscillator. Characteristics of these tubes at 18 MHz are given in **table 1** with those of some replacement semiconductors. Problems occurred with rf gain, rather than with interelectrode capacities, because of signals feeding back along the tuning capacitor rotor (from gang 3 to the antenna section, for instance) before other effects showed up. Also, the 3N140 will oscillate enthusiastically at any frequency up to 300 MHz as I found out.

The extra resistors and capacitors in the circuit are mainly to kill various sorts of vhf parasitics. In order to get the rf gain down where it belonged (high rf gain will also cause difficulty with mixer overloading) it was necessary to put capacity voltage dividers

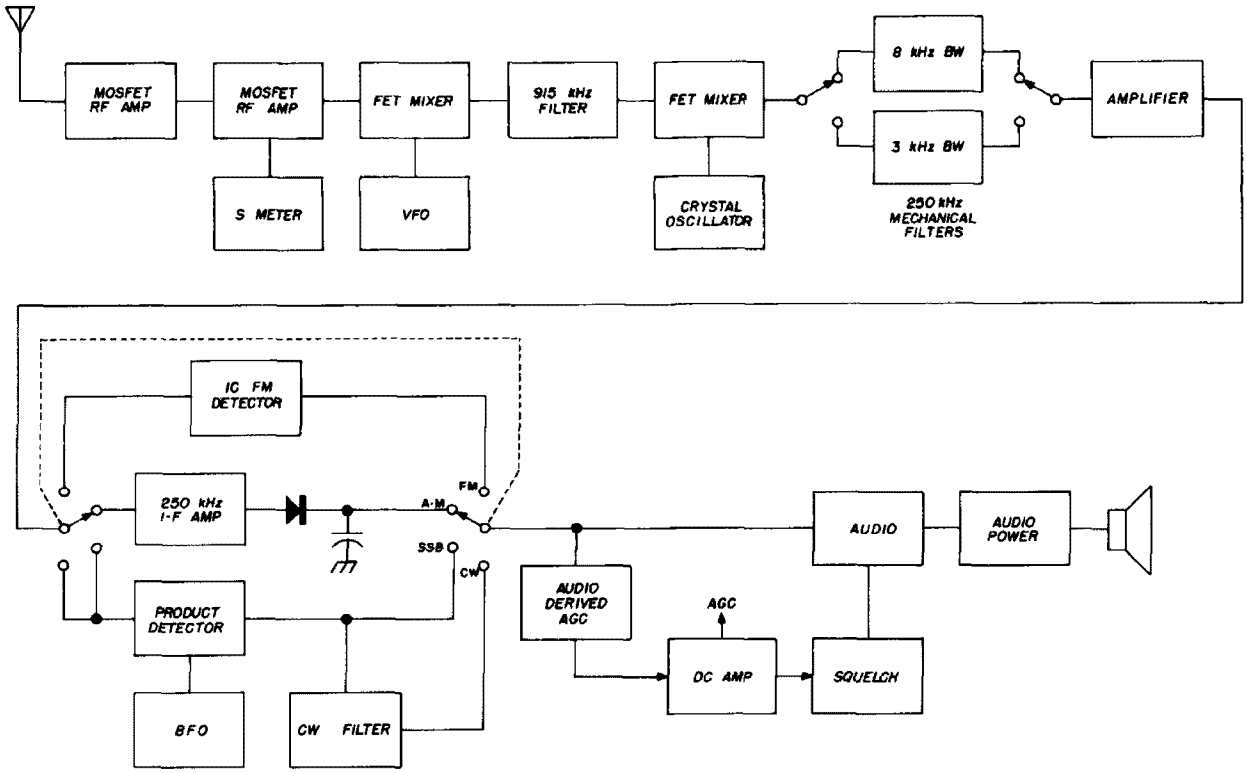


fig. 1. Simplified block diagram of the solid state BC-348 receiver. In the actual receiver, the audio-derived agc circuit is switched out of the system in the a-m mode.

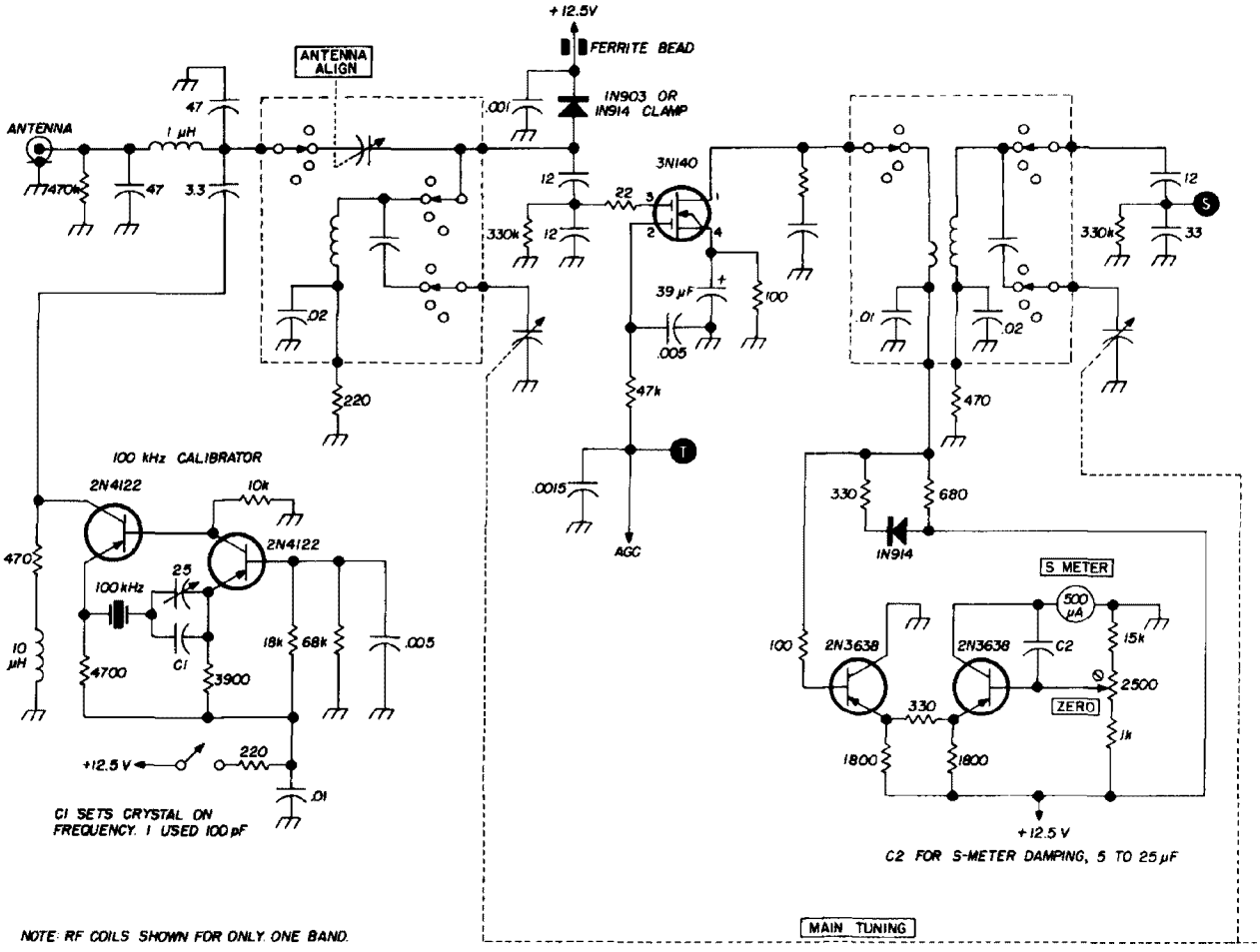


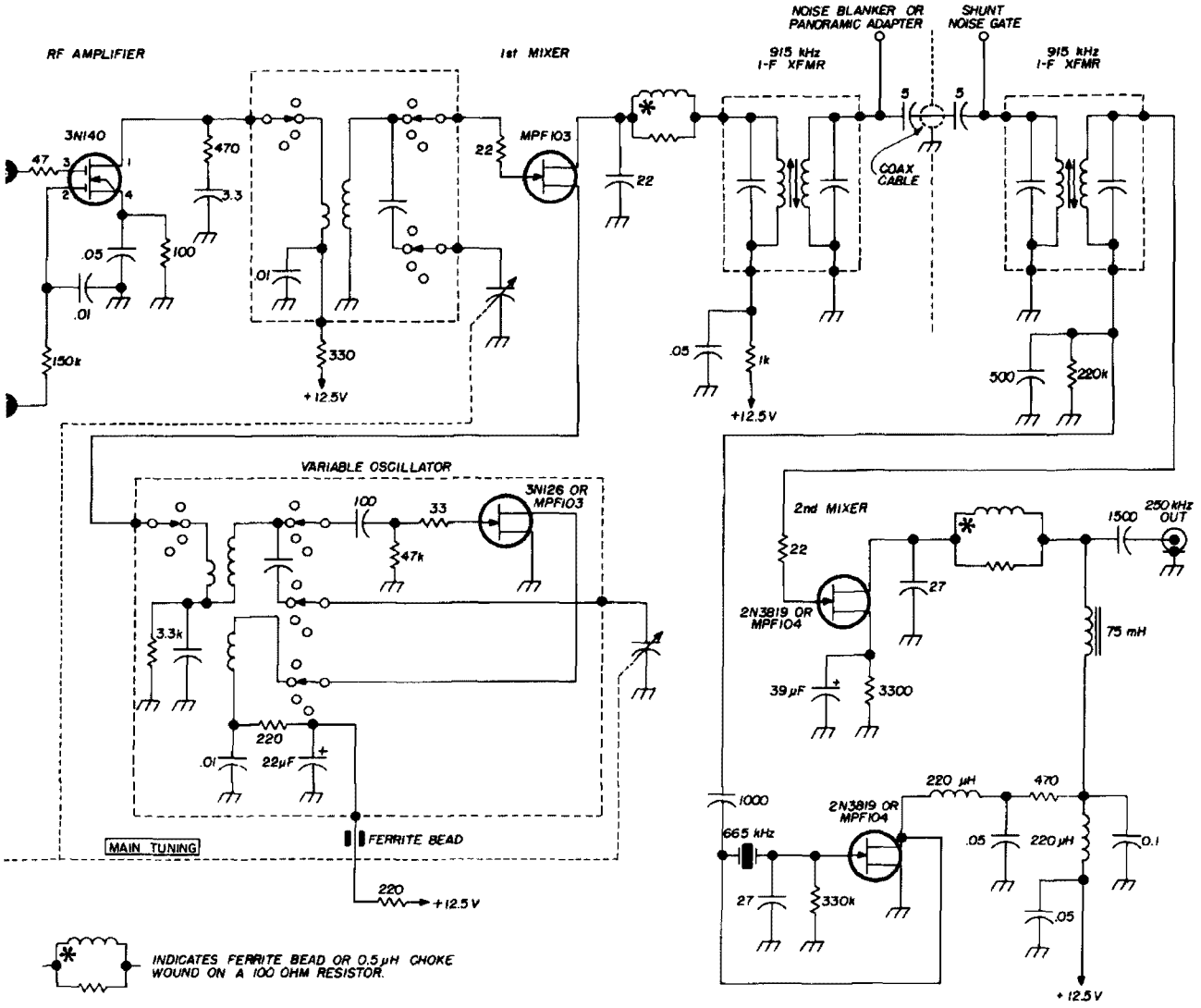
fig. 2. Schematic diagram of the solid-state BC-348. Coils in the rf section are shown for one band only; band-switch shaft end should be grounded with a spring to remove 144-MHz birdies. Resistors are ¼-watt composition unless otherwise marked; 1% resistors are 1/8-watt metal-film types (RN55 or RN60). Polarized capacitors less than 25-μF are tantalum types (150D or CS13); aluminum electrolytics are ok if not used for agc or audio coupling.

at the input of each 3N140. Because the semiconductor triode noise is lower than that of the 6K7, the measured noise figure of the modified receiver was somewhat better than the original, even though there wasn't always a peak in noise as the antenna trimmer was turned.

Gain control is provided by changing the bias on the second gate. Insulated gates don't draw current, but the control voltage should be +4 V for maximum gain, and complete cutoff is obtained at −3 V. The first gate is returned to ground. There's no chance of rectified current causing the whole receiver to block when the rf stage conducts (a common cause of receiver overload).

A triode FET can replace the 6J7 mixer. The high feedback capacitance is a minor problem so long as the input and output circuits are tuned to widely different frequencies. The conversion gain is a function of the available oscillator drive power. With enough injection the mixer gain could be ten times that of the original. To replace the 6C5 oscillator a low-transconductance triode FET is needed, as discussed below.

I went to double conversion to obtain better selectivity. If 455-kHz filters were to have been used, I would have moved the first i-f to 930 kHz or higher in order to avoid two intermediate frequencies in harmonic relationship. Of course the second oscillator would have been put on the high side. (Think where the image would be if it were on the low side.) However, I had some 250-kHz mechanical filters on hand. The point of this conversion was, after all, to use the six-band, twenty-four-coil tuning appara-



tus of the original receiver with as little modifications as possible.

Fig. 1 is a block diagram of the complete receiver, including auxiliary features (to be covered in another article). Everything fits inside easily. **Fig. 2** is a schematic of the main rf and i-f circuits.

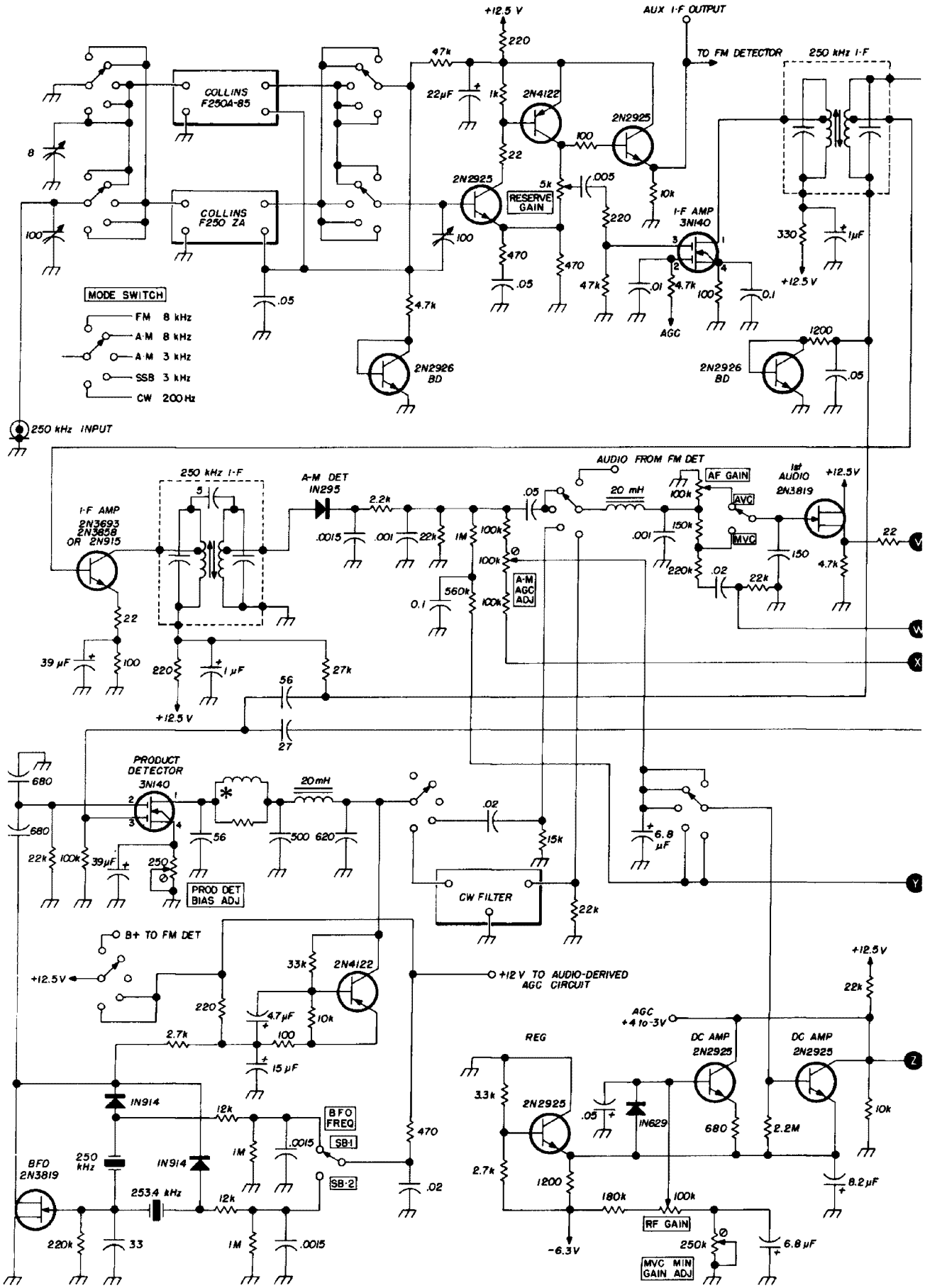
input circuit

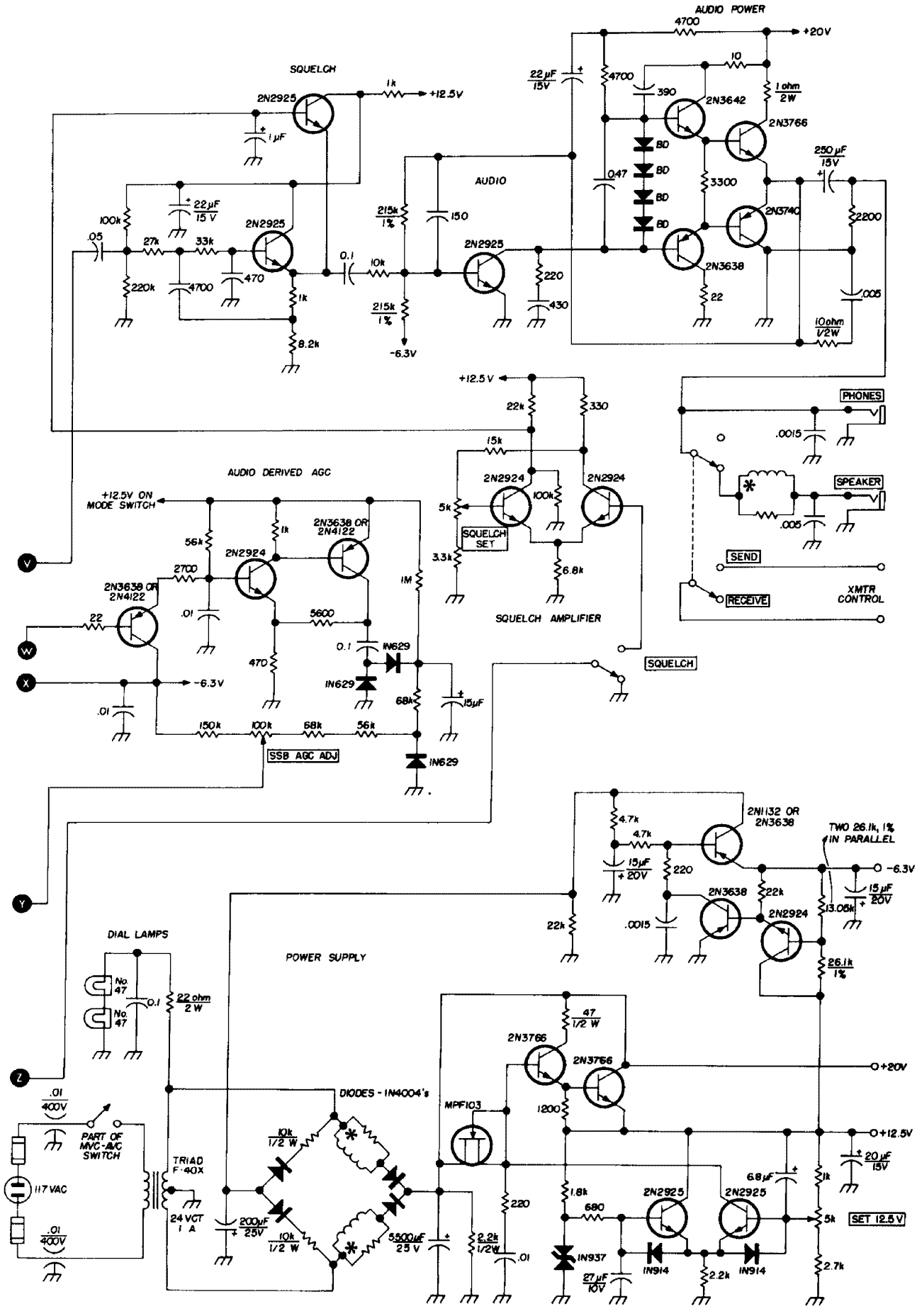
The antenna input starts out with a low-pass filter, with nominal cutoff at about 30 MHz. Aside from obvious advantages if you are near strong TV stations (I am), there are also less "birdies" when the receiver is used with a vhf converter. Because of the input coupling method in this model receiver, the filter is especially important.

There is a protective diode clamp on the antenna coil. 6BA6 rf amplifiers will stand several hundred volts swing without damage. If the grid return resistance is high, however, the diode is essential because the insulated gate FET is rated for only a few tens of volts peak. A type 1N914 of any manufacture will have low loss at 12 volts of reverse bias. The added capacitance is about 1 pF.

The change of drain (plate) current in the first rf stage is used to indicate signal strength. Because of the slope of the 3N140 control characteristic, the meter scale is nonlinear in dB, even with the diode in the circuit. Other approaches, including a separate signal indicator channel, were less satisfactory.*

* The use of an s-meter as a quantitative measuring device for received signals is open to question, but an s-meter is fine as a tuning indicator. **Editor**


+12.5 V
AUX I-F OUTPUT
TO FM DETECTOR
250 kHz I-F
COLLINS F250A-85
COLLINS F250 ZA
MODE SWITCH
FM 8 kHz
A-M 8 kHz
A-M 3 kHz
SSB 3 kHz
CW 200 Hz
2N4122
2N2925
RESERVE GAIN
I-F AMP 3N140
2N2926 BD
AGC
250 kHz INPUT
I-F AMP 2N3693 2N3658 OR 2N915
A-M DET IN295
AUDIO FROM FM DET
AF GAIN
AVC
MVC
1st AUDIO 2N3819
A-M AGC ADJ
PRODUCT DETECTOR 3N140
PROD DET BIAS ADJ
CW FILTER
B+ TO FM DET
+12 V TO AUDIO-DERIVED AGC CIRCUIT
REG
DC AMP 2N2925
IN914
IN629
BFO 2N3819
BFO FREQ
SB-1
SB-2
253.4 kHz
RF GAIN
MVC MIN GAIN ADJ
-6.3V


AUDIO POWER
SQUELCH
AUDIO
AUDIO DERIVED AGC
SQUELCH AMPLIFIER
SQUELCH SET
SSB AGC ADJ
SQUELCH
PHONES
SPEAKER
SEND
RECEIVE
XMTR CONTROL
DIAL LAMPS
POWER SUPPLY
DIODES - IN4004's
PART OF MVC-AVC SWITCH
TRIAD F-40X
117 VAC
24 VCT 1 A
TWO 26.1k, 1% IN PARALLEL
SET 12.5V
+20V
+12.5V
-6.3V

local oscillator

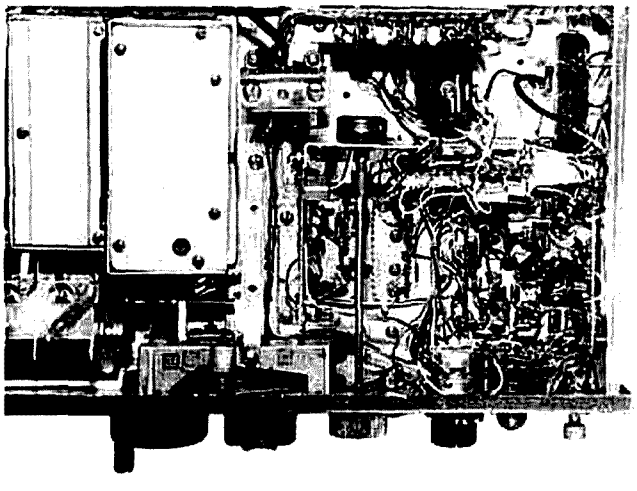
The local oscillator FET socket is supported by its leads, which pass through the keyway from the octal socket pins. The 6C5 ran at around 90 volts, with a starting g_m of less than 2 millimhos. I tried various adaptations, and found that "hot" FET's (those having high I_{dss}* and a high zero-bias g_m) were noisy and unstable, especially at about 15 MHz.

Good results as a local oscillator were obtained with the Motorola 3N126 and some MPF103's. The 3N126 was used with the second gate (substrate and case) hooked to source. If a MOSFET is used, something like a 1N34 would be needed to generate "grid-leak" bias.[1]

In the original oscillator the grid leak was 100k ohms. In the course of chasing troubles I changed the value to 47k, but 100k probably would be satisfactory. The 15k mixer cathode resistor, also inside that box, probably would be satisfactory without change, although mixer FET's having very high I_{dss} would be more suitable in that case. The 3.3k resistor I used was based on experience with the breadboard model of the second converter. The 33- (or 47-ohm) resistor in series with the gate lead of the oscillator can be put in between the octal socket (pin 5) and the transistor socket.

mixer

The mixer proved to be rather critical as to the FET that was used. For type MPF103 the I_{dss} range is 1 to 5 mA; for the 2N3819 it is 2 to 20 mA. I found that anything between 1 and 4 mA worked in my receiver, but the higher-current units gave a bit more gain. Still higher I_{dss} units had low gain and were noisy in this socket. Other places in the receiver are not so finicky in this respect. I'm not worrying about spares: if the FET's last as long as 6SK7's, I'll need another one in 1990.



In a double-conversion receiver, there should be as much selectivity at the first i-f as other considerations permit, and as little gain between the first and second mixer as possible without compromising noise figure (but in no case less than unity gain, or the first stage will overload before the second). A few calculations show that three very high-Q tuned circuits will reject the nearest spurious response (789 kHz), but four tuned circuits of moderate Q will do even better.

I could have used two of the original cans, but I rewound a couple of miniature J-trans to save space. (For details on how to adjust the coupling using a Q-meter, see reference 2.) The "tee" of capacitors between the cans is used in setting up the gain, as the coupling factor between pairs is much less than critical. The shunt capacitance is that of the coax running from the rf board to the second converter module.

second conversion and detector stages

The second mixer and oscillator are built on a small plate that fits in place of the original output transformer. The values were juggled to get best operation with the FET's and crystal I had. With my setup, the oscillator drain swing was about 9 V p-p, and the mixer developed no rectified bias across the 220k resistor.

Strong second oscillator harmonics were found at the mixer output and on the B-supply lead beyond the feed-through capacitor. The additional rf choke and capacitor fixed the power-lead leakage, while the natural attenuation of the mechanical filter is enough protection for the output harmonics,

* I_{dss} is a common classification parameter, measured by shorting the gate to source and applying 5 to 10 V to the drain. Thus by definition, I_{dss} is the drain-to-source current when the gate-to-source voltage is zero.

table 1. comparison of tube and semiconductor characteristics.

Type		V_B (volts)	I_B (mA)	g_M (millimho)	C_{FBK} (pF)	R_{in} at 18 MHz	R_{out} (ohms)	R_{eq} (noise) (ohms)
6C5	tube	80	6	2	2	50k	10k	—
8J7	tube	100	2	1	0.005	60k	800k	10k
6K7	tube	100	9.5	1.6	0.005	50k	150k	15k
2N2084	padt	12	1	34	2.0	800	40k	200
3N126	jfet	12	2	1.4	0.5	25k	30k	—
3N128	mosfet	15	2	4	0.2	100k	80k	—
3N140	mosfet	12	10	8	0.02	100k	12k	<1k
MPF103	jfet	12	2	3	2	100k	50k	<2k
40481	npn	7.5	2	84	0.2	1500	50k	200

if the filter input terminals, etc., are well shielded.

Second mixer output is parallel-fed via a choke and blocking capacitor to the filter switch. Either the 3-kHz or the 8-kHz bandpass filter may be selected, the other one having its terminals shorted.

The ouput section of the switch (8 pole, 5 position) goes to a feedback amplifier that has a high-impedance input and a gain of ten, followed by a potentiometer that permits the background noise level to be adjusted without modifying the agc action. The range of the control is 23 dB.

The stage following the gain control is the third 3N140. It is coupled by an auto radio i-f can, padded down from 262 to 250 kHz with additional 22-pF capacitors, to the product detector and to an i-f power amplifier that drives the diode detector. The two i-f cans shape the top of the passband in the 8-kHz bandwidth a-m position to get the best weak-signal a-m reception. (This seems to want a rounded top, symmetrical, down about 6 to 10 dB at 3-kHz off center.)

The front-panel switch selects five combinations of bandwidth and detection, as per **table 2.** The agc source for a-m is the diode detector, but for sideband and CW reception agc is derived from rectified audio, so that the effect of audio selectivity is included.

mode-selection circuit

The mode switch is shown for 8-kHz bandwidth a-m reception, with agc in operation. The second-gate control voltage on the three 3N140's varies from plus three or four volts with no signal to negative two or three on very strong signals. The loop gain is high, giving a dynamic range of 40 dB. The first volt of bias change makes very little change in receiver gain, but the squelch operates reliably at that point.

agc/mvc

In the gain control circuit the two dc amplifiers, one to handle agc and one for manual control, are hooked to a transistor that supplies —4 V to the emitter returns and which, incidentally, provides temperature compensation—a sort of three-transistor differential amplifier.

A negative voltage supply makes the am-

table 2. Mode switch setup.

function	mode FM-8	AM-8	AM-3	SSB	CW	REMARKS
filter (kHz)	8.5	8.5	3	3	3	4 - poles
200-Hz audio filter	no	no	no	no	yes	1 - pole
age	car	car	car	af	af	1 - pole
B+ to SSB and bfo	off	off	off	on	on	
audio from	fm	a-m	a-m	prod	prod	1 - pole
B+ to fm dot	on	off	off	off	off	1 - pole for B+

notes:

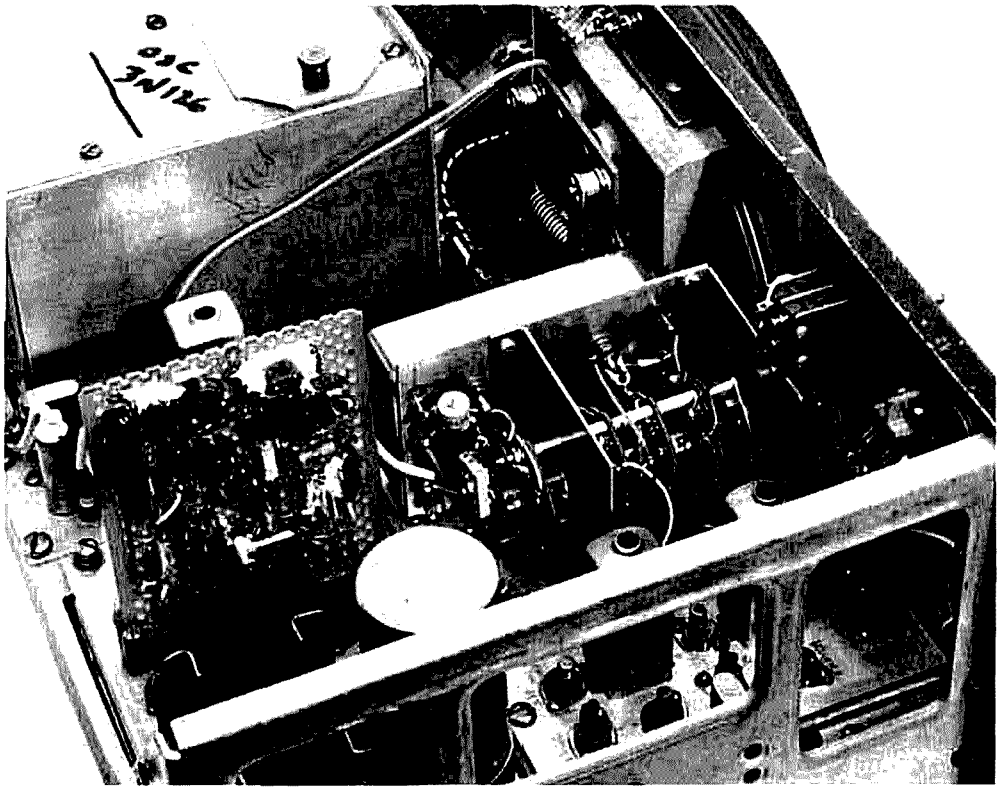
1. SB1-SB2 (SB1 is lower 40-75, upper, 20).
2. Agc-off-mvc functions similar to original.

plifier practical and is a handy way to make dc level changes without extra stages. Also the audio stage operating point is centered automatically.

The audio-derived agc voltage is from a little three-transistor feed-back amplifier (with automatic temperature compensation; again, this is from the pnp follower driving the npn stage) and a voltage-doubling rectifier. The attack time is fast for small changes, but the maximum effect from a single pulse is limited by a 0.1-microfarad capacitor that couples the amplifier to the rectifier, while the rectifier has fifteen microfarads to charge up.

For 1-kHz CW tones, this means that 1 Hz makes a maximum difference in charge of the 15-μF capacitor of 10 V times 0.1 divided by 15, or 0.067 V. However, the gain of the dc amplifier after that is considerable, and gain changes of 20 to 40 dB/millisecond can be obtained. If action were much slower, there would be worse thumping at the beginning of a sideband transmission. If the action is too fast, there may be a stability problem or there may be a tendency for the receiver to block on a burst of QRM or ignition noise.

For fm reception, a Sprague integrated circuit is used, type ULN-2111A. This has a bunch of emitter-coupled clipping amplifiers, a rather fancy six-transistor phase detector, and various diode bias networks. All that is required outside is a single-tuned circuit. (A typical transistor radio i-f can is suitable; just use the high impedance winding, padded to frequency with external capacitors and some by-pass capacitors.) The output in my receiver swung three volts peak-to-peak for a signal moving across the 8-kHz filter. When the signal was injected after the filter, an s-shaped curve about 18-kHz wide was obtained, with an amplitude of more than five volts peak-to-peak.



Chassis view of the revamped BC-348 shows the new mode switch and two mechanical filters. The power supply and audio circuits occupy the old dynamotor space; the plug-in board contains voltage regulating and low-level audio circuits.

For manual gain control (I happen to like to dive for the same volume control knob whether agc is on or not), the agc circuits are still operative to prevent blocking on unexpectedly strong signals. The audio gain is high enough so that normal operation is in a region where the agc threshold is not exceeded. Because of the highly nonlinear control function (dB versus control volts), tapered sections are used for both audio and manual rf gain.

diodes. However, 1N914's could be used in place of the 1N457's as well.

The 2N2925's and 2N2924's were actually the green- or yellow-coded 2N2926's, while the brown- and red-coded 2926's were used for bias compensation diodes, labeled "BD." You get 25 assorted for $7.25.** You'll be able to use 20 if you're crazy enough to copy this receiver.

There are also two 2N3638's, three 2N4122's (two in the calibrator, which is

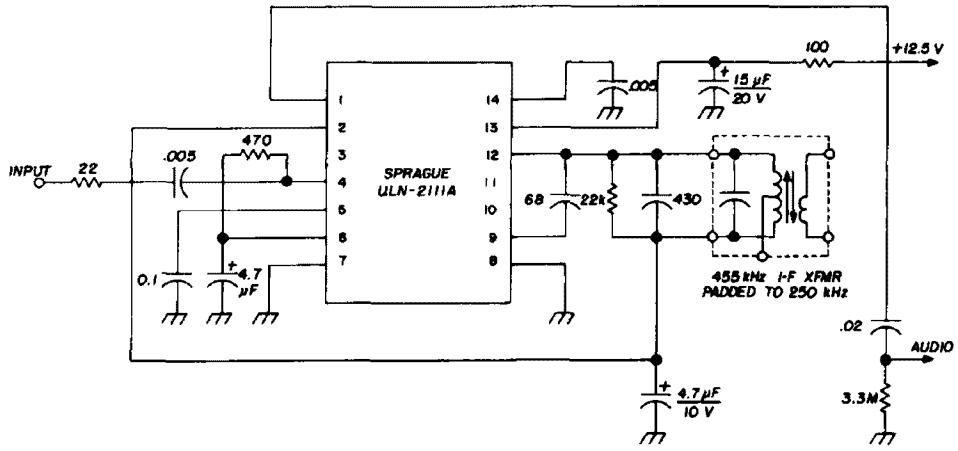


fig. 3. Integrated-circuit fm detector using a Sprague ULN-2111A. This circuit provides 3 volts p-p output with 1 millivolt input.

I found that a dual 100k audio taper unit was satisfactory; the value could be as high as 250k and as low as 50k with a few minor adjustments. A stereo amplifier control should be adequate. The **reserve gain** pot in the 250-kHz i-f amplifier could have either a linear or audio taper such as used in transistor radios.

The rest of the controls (for agc thresholds, product detector bias, manual volume control "off" bias,* and the power supply voltage setting) can be any handy trim pot, either surplus from computer cards, or small trimmers such as the IRC 2CI or Mallory MTC.

bfo circuit

High-speed computer diodes such as the 1N903 or 1N914 are used for the input voltage clamp and the bfo crystal switch; I used 1N457's or 1N629's for other spots, as I had them, and I felt that there was less chance of generating harmonics with the "slow"

mounted on the rf panel near the input) and six pnp's, which could be either type. The first type costs fifty cents; the second eighty. (The whole count is 50 transistors and FET's, 16 various diodes, and one integrated circuit.)

audio section

After the mode of reception is selected (with or without narrowband CW filter), the signal goes, via the audio gain pot in the agc position, to a source-follower first audio stage, an active low-pass filter, then to a complementary class-B output stage. The squelch is in the active filter section. The circuit is

* Minimum gain setting must be at minimum gain bias.

** Available from Allied Radio Corporation, 100 N. Western Avenue, Chicago, Illinois 60680. Order stock number 49F3-2N2926. Kit of 25 is $7.25 plus postage.

similar to one previously described.[3]

The power regulator, power transformer (Triad F-40X 26 V, 1 A center-tapped), squelch control, and all except the first audio are mounted in the space vacated by the dynamotor. The high-power transistors are on the subchassis (insulated by mica washers), and the low-power circuits are on a plug-in card (very easy to pull out and change).

selectivity filter

The CW filter is switched into the sideband channel for more selectivity when needed. The values seem reasonably satisfactory in practice, although the filter performance has not been measured in place in the receiver. Any other narrowband audio filter of ten or twenty thousand ohms impedance should do as well. Obviously, an extra amplifier stage could be put in the chain if the filter on hand was of lower impedance, as the method of switching makes this practical. The audio agc is based on what passes through the filter, so it can be used even under quite rough conditions. Because the squelch can also be used in this mode, I have been able to find a fairly weak (though **stable**) signal and park on him with the receiver on squelch, waiting to get a chance to work him.

crystal calibrator

The crystal calibrator is almost a necessity in a wide-range unit such as the BC-348. It appears that there would be some benefit in making a more complex calibrator, such as a 200-50-10 kHz type. Higher output than I have is desirable when running some converters, but I haven't done anything about it as yet.

in retrospect

The main problems with transistors almost always stem from overloading and temperature variation. Resolving the overload problem takes effort: care to see that the last i-f amplifier, for instance, is capable of driving the second detector hard enough, so that the second detector can generate enough agc voltage, so that it won't be driven too hard—like a tight servo system. (The more I read that the more I wonder if the English language is up to today's problems.)

What all this means is that the last i-f stage should be designed for best power output, not best gain. Also, it means that agc can't be used on the last i-f stage. (Auto radios with **one** i-f stage have some rather fancy solutions to this problem.)

Another device to reduce overload is feedback. At "low" frequencies (under a couple MHz for today's transistors), feedback will hold gain and bias constant despite temperature variation. Distortion will also be kept low.

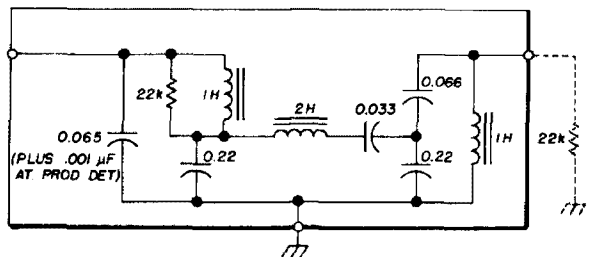


fig. 4. Audio CW filter with nominal 700-Hz center frequency and 160-Hz bandwidth. Inductors are Triad epoxy-molded toroids with Q of 50 to 80 at 1000 Hz. Capacitors are mylar insulated.

In many places where dc levels are critical, differential amplifiers are used. The drift with temperature of one transistor is balanced almost exactly by another of the same type. In other circuits, a diode-connected transistor (or any diode) is used as part of the dc bias circuit, so that temperature effects of the transistor are pretty well cancelled by the diode as it conducts a few mA in the forward direction. In class B or AB circuits this is particularly important. For best results, the bias diodes must be placed physically so that they are at about the same temperature as the power transistors.

references

1. G. D. Hanchett, "Insulated Gate Field Effect Transistors in Oscillator Circuits," *RCA publication ST-3520.* (Available from P. O. Box 53, Harrison, N. J.)
2. F. Langford-Smith, *"Radiotron Designer's Handbook,"* 1952 edition, page 1030.
3. H. H. Cross, W1OOP, "An Obsolete Mobile Receiver," *QST*, November, 1967, p. 11, and December, 1967, p. 31.

ham radio

power supplies for single sideband

Forest H. Belt, 2610 Whittier Avenue, Louisville, Kentucky 40205

A power supply's a power supply. Or, so it seems. But not all of them suit all purposes. Sometimes extra demands are made on power supplies for single-sideband equipment—such as wider regulation, multiple voltages, cleaner filtering, and so on. That's why these dc supplies deserve special attention.

First, though, it's just as well to review the characteristics common to all power supplies. What are they supposed to do, and how do they go about it?

Mainly, the power supplies in ham equipment furnish dc voltages to operate tubes and transistors in the stages. Circuits may need positive voltages or negative voltages, high voltages or low voltages, high currents or low currents. The power supply generally converts 117-volt ac power to the needed dc voltages.

A power supply also furnishes certain specific ac voltages. The power transformer usually develops them with an extra winding or two; you don't even notice them. Without them, though, tubes wouldn't get proper heater voltages—and sometimes relays wouldn't work.

getting the voltage you want

High-power ssb transmitters naturally need much higher voltages than those needed by, say, a receiver. Transistor equipment needs lower voltages. There are three chief methods for getting the exact voltage needed by each stage of a transmitter or receiver.

The first way is by transformer. By specifying transformer windings with certain turns ratios, the ssb-equipment designer gets a range of voltages. After the ac voltages are at the right values, they can be changed to dc.

A step-down turns ratio of 117:6.3 supplies the ac voltage for most parallel-wired tube heaters. If a special transmitter tube needs 5 volts or 10 volts for its filament or heater, a winding ratio of 117:5 or 117:10 is used. For higher voltages, set-up ratios are used. For 350 volts, a turns ratio of 117:350. For 1000 volts, a ratio of 117:1000.

As the transformer schematic in **fig. 1A** suggests, any reasonable number of windings can be included in a single housing. Also, as **fig. 1B** illustrates, one winding can be tapped to provide several voltages. The transformer method is expensive, but it's a practical way to step the power-line voltage either up or down.

The second way to get specific dc voltages from a power supply is by **resistive divider.** This method is used after the volt-

age has been changed from ac to dc. A drawback is that you can only step voltages down—not up.

The power transformer steps the power-line ac up to the highest voltage desired. After the ac is changed to dc, the dc voltage is fed to a series of resistors; the voltage divides across the resistors, in proportion to the resistance values. An example of this appears in **fig. 1C.**

The third way to get the voltage you want is by **voltage doubling.** This is a trick that builds voltages up in value, but can't step them down. Voltage doubling is done in the rectifier part of the power supply. The result is a dc voltage that's twice the value an ordinary rectifier would produce. Where especially high voltages are needed, tripling and quadrupling can be arranged.

In commercial ssb equipment, you'll find combinations of these three ways. Each has certain advantages, and in combinations they are versatile. You should consider all of them for homebrew ssb gear.

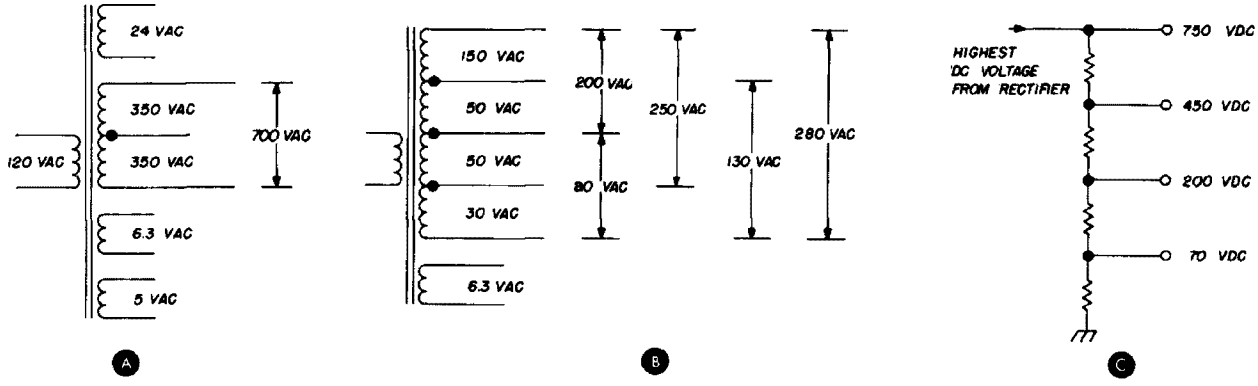


fig. 1. There are several ways to get the voltage you need; by selecting the power transformer, by using a tapped secondary or with a resistive voltage divider.

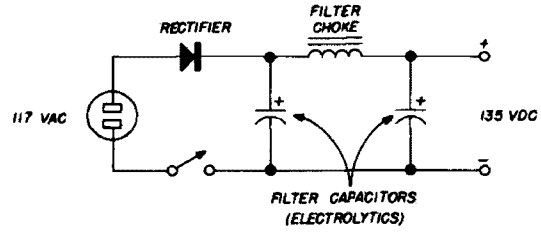
from ac to dc

Ac power is changed to dc power by rectifying and filtering. A tube or semiconductor is the rectifier. Nowadays, semiconductors are more popular. The rectifier conducts on half of each ac cycle and blocks on the other half. The output is pulsating dc—which is dc voltage with a rather large ac component still riding along. Before the dc is fit to use, the ac must be removed; electrolytic capacitors and filter chokes do this. The output of a rectifier-filter combination is relatively pure dc. A simple example appears in **fig. 2.**

The dc output voltage depends mostly on how much ac is applied to the rectifier. The line voltage can be applied directly, as it is in **fig. 2,** or, ac can be applied from a stepup winding of the power transformer to make the dc output voltage greater. Ac can also be applied from a stepdown winding, as it is to develop low-voltage dc for powering transistors.

The dc output is usually slightly higher than the measured ac voltage applied to the rectifier. The output of **fig. 2,** for example, is about 135 volts dc, although the input is only 117 volts dc. That's because the output depends on the peak value of each cycle, whereas the measured ac voltage is rms or effective value.

fig. 2. Simple half-wave rectifier circuit with capacitive input filter.



Rectifiers are broadly categorized as half-wave and full-wave. A half-wave circuit is shown in **fig. 3A.** Half-wave rectifiers waste one half-cycle of the input waveform. Both halves are applied, but only one can pass. In **fig. 3A** it is the positive half, because of the direction the rectifier diode is connected. The filter capacitors and choke smooth out the deep ripples, leaving a positive dc voltage with almost no ac in it.

A full-wave rectifier is shown in **fig. 3B.** It requires a center-tapped transformer winding; the center tap is grounded in this version.

Considering the whole secondary winding, each positive half-cycle makes the top rectifier conduct and each negative half-cycle makes the bottom one conduct. With respect to ground, however, both rectifiers receive a positive half-cycle. The combined output of the two rectifiers is a series of positive half-cycles. The full input wave is used, instead of just half of it. The filters make the output nearly pure dc.

A bridge rectifier circuit is shown in **fig. 4.** This one is a full-wave system because it uses both half-cycles. Compare this transformer with the one in **fig. 3B,** and you can see why the bridge circuit is popular. The 280 volts dc output is obtained with a simple 250-volt transformer instead of a center-tapped 500-volt unit.

One more circuit that converts ac to dc is the doubler. There are two common kinds, both shown in **fig. 5.** The one in **fig. 5A** is a half-wave doubler. Consider the bottom of the secondary as a reference point. The first negative-going half-cycle charges C1 through D1; D2 can't conduct because the voltage is in wrong polarity. Then, the positive half-cycle that follows charges C2 through D2. The previously stored charge on C1 is in series, and it adds to the positive half-cycle. The charge that develops across C2 is therefore double what it would otherwise be. This doubled charge is applied to C2 with each positive half-cycle; the negative half-cycle doesn't affect C2 at all—only C1. The output is therefore a half-wave pulsating dc voltage almost double the peak input ac voltage.

The full-wave doubler in **fig. 5B** uses both half-cycles. Each positive half-cycle charges C2, because D2 can conduct. The negative half-cycle that follows then charges C1, because D1 can conduct. Each half-cycle of the input wave therefore charges one capacitor or the other. The output is taken across the two capacitors in series, so their voltages are added. The output is a full-wave pulsating dc voltage that's about double the peak input ac voltage.

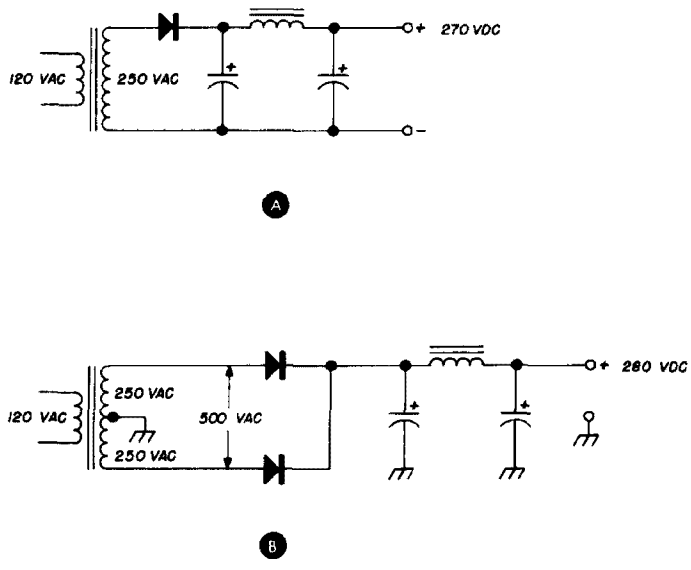


fig. 3. Two types of rectifier circuits—half wave in A and full wave in B.

the current you need

Transistors don't require much dc current in most uses. Tubes used as small-

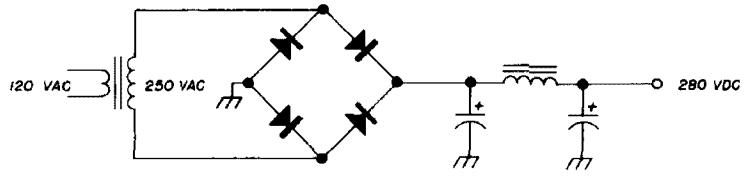


fig. 4. The full-wave bridge rectifier.

signal amplifiers need more; tubes for receiver power amplifiers need still more; tubes in transmitters need even more; and transmitter power tubes need plenty. Single-sideband linear power amplifiers don't use as much average current as their a-m or cw counterparts, but their demands during modulation peaks are even greater.

The range of current that single-sideband power supplies must furnish is wide, then. Three factors about a dc power supply determine its ability to deliver current without overheating or damage: the transformer-winding current rating, the rectifier current rating and the size of the filter capacitors.

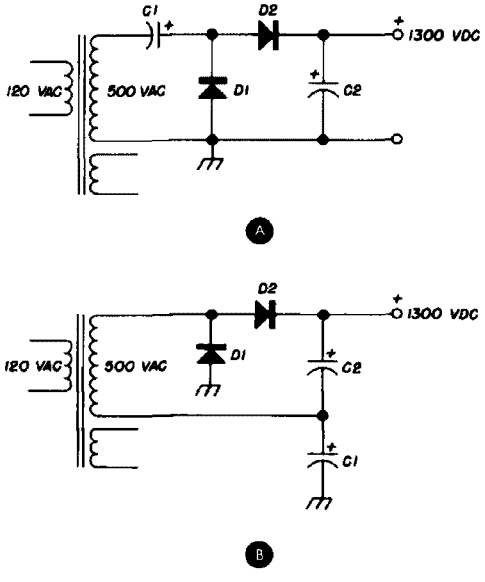


fig. 5. Here are two circuits which rectify the ac voltage and double it at the same time.

When a designer knows how much current all the tube plates and screens draw, he picks a rectifier (or group of them) that can handle the current. The rectifier must also be able to withstand the voltage that is to be applied. (Rectifier diodes can be wired in series to increase voltage ratings.) Then the designer picks a transformer that supplies enough ac voltage to develop the maximum dc voltage needed and makes sure its windings are rated to carry the needed current.

Filter capacitors store electrons that will be drawn from the power supply as current. If the filters are small and the current drain is heavy, the pulsations of dc don't have time to "fill them up;" the current keeps being drawn out too rapidly. Larger capacitors store more electrons, and current demands don't deplete them between half-cycle pulsations.

Full-wave rectifier circuits supply more current than half-wave circuits. Their pulsations are twice as frequent as those of half-wave rectifiers, so they keep the filters charged up better.

Voltage doublers reduce the current a rectifier can supply. Double the voltage, and you halve the current that can be drawn safely. From a tripler, only a third as much current is available as from the same set of components in a simple rectifier and filter.

smoothing it out

Getting rid of whatever ac is still a part of pulsating dc is the job of the filter capacitors and choke. The larger the capacitance is, the less ac gets through. Also, the larger the choke inductance, the smoother the dc that results.

Some typical pi-network power-supply filter circuits are shown in **fig. 6.** The input filter (C1) has a lot to do with the output voltage. The larger it is, the higher the dc —up to nearly the peak value of the input ac. The output filter controls how much current the supply can furnish without leaving a lot of ac ripple.

Aside from their effects on voltage and current, the two capacitors cut down ripple. They are low impedances for power-supply ac. The input capacitor "shorts" a lot of the leftover ac (from the rectifier) to ground. The choke has a high impedance to ac but passes the dc easily if it's rated to carry enough current. The output capacitor then "shorts out" any remaining ripple.

When not too much current is to be drawn, a wirewound resistor can take the place of the choke. **Fig. 6B** shows the circuit. The resistor isn't as effective as a choke, but it works.

The diagram in **fig. 6C** shows the elec-

trolytic capacitors "upside down." That's because the input is negative pulsating dc. Electrolytics are polarized and must be connected correctly. The output is smooth negative dc.

dc plus or minus

The last power-supply characteristic we'll review is polarity. In tube-type ssb receivers, the voltages needed most are for plates and screens; the supplies therefore supply positive voltages. Transmitters and linear amps sometimes need high negative bias voltages, so the power supply must develop a negative output voltage. Transistors, too, may demand negative supply voltages—usually much lower than those for biasing tubes.

Polarity of the dc output from a power supply is determined by the direction in which the rectifier is hooked up. If the cathode (bar) is on the side toward the output or the filter network, the output is positive. If the anode is in that direction, the output is negative. The sketches in **fig. 7A** show both.

There's another way to get negative voltages, without using a separate rectifier. It's diagramed in **fig. 7B.** This is an ordinary positive-output dc supply, furnishing 105 volts dc across the output filter. The bottom of the transformer secondary is not grounded; instead, the ground point is between the two resistors that form a bleeder across the output.

Dc voltages for stages in modern equipment are measured with respect to ground. A voltmeter connected to the bottom of the bleeder in **fig. 7B** measures −5 volts dc. Thus, by the simple trick of moving the ground point up the bleeder chain, a negative voltage becomes available. In ssb equipment that requires both positive and negative supply voltages, this is a favorite way to get them.

ssb power supplies at work

The power supply in **fig. 8,** chosen from an ssb receiver, demonstrates many of the principles just discussed. The main supply

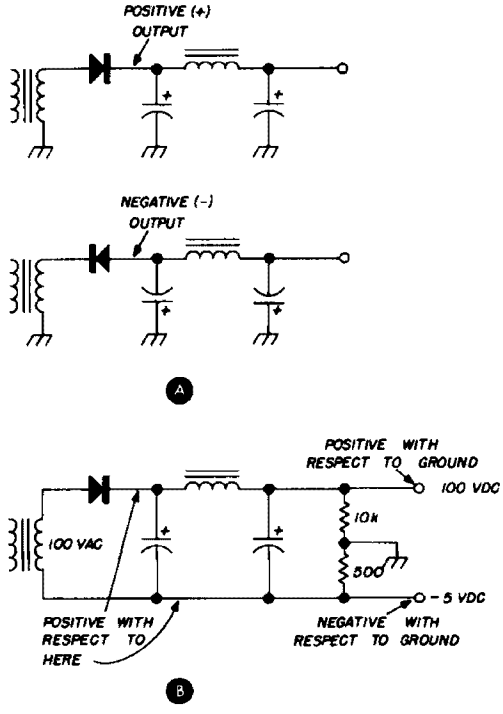
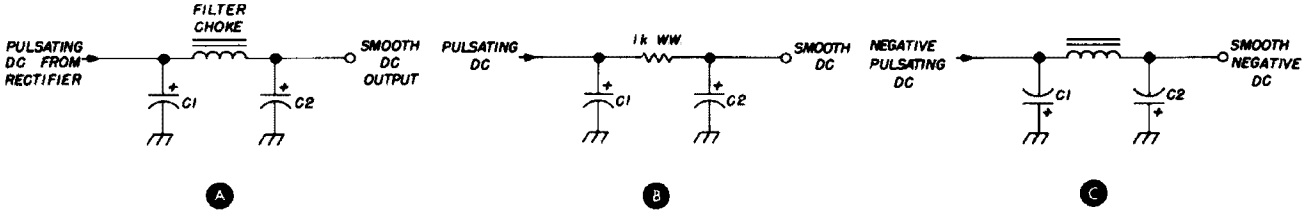


fig. 7. Positive and negative voltages with a single rectifier.

is a full-wave rectifier. D1 and D2 supply positive-going pulsations to a large input filter. With 180 volts ac from each transformer winding, dc across the output filter is 195 volts.

Some circuits in this receiver need closely controlled voltage. So, a regulator tube holds the output at the end of the 4k re-

fig. 6. Pi-network power-supply filters.



sistor at exactly 150 volts—no matter how much current is drawn from that branch of the supply (up to 35 mA).

Also, a negative voltage is needed. For it, the lower half of the secondary winding applies 180 volts ac to a resistive voltage divider, lowering the ac voltage applied to

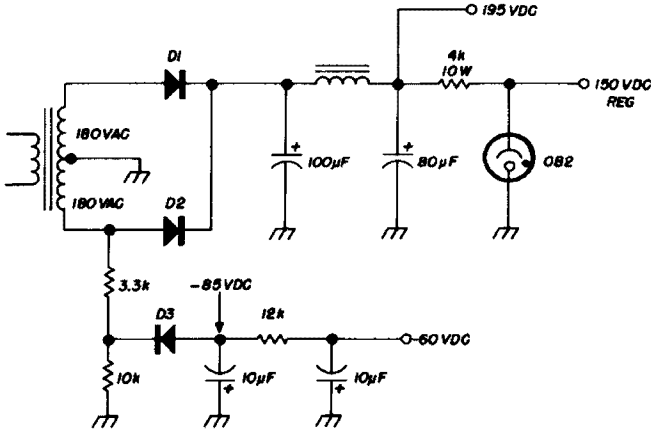


fig. 8. Power supply for an ssb receiver demonstrates many of the power-supply principles discussed in the test.

D3. The rectifier is connected for anode-output, so it supplies negative pulsations to its filter network. The capacitors, connected for negative output polarity, work with the 12k resistor as a pi-network filter to smooth out any ripple. The output is −60 volts dc.

One transmitter supply uses some interesting variations. Take a look at **fig. 9.** The arrangement is part full-wave doubler and part ordinary half-wave rectifier. The rectifier diodes are in series merely to increase their voltage rating, so less-expensive diodes can be used.

First, the half-wave circuit. It consists of D3, D4, the full transformer winding, and filter C2. When the top of the transformer winding is negative with respect to the bottom, D3 and D4 can conduct, charging C2. On the next half-cycle, when the top is positive-going, they can't conduct. No more voltage is added to C2, nor can it discharge through the rectifiers. Thus, a train of positive half-cycles applied to C2 develops a positive pulsating dc voltage there. The capacitor is large enough that the pulsations are mostly smoothed out. The output is about 260 volts of fairly smooth dc.

The 200-ohm resistor and C3 form a pi-network with C2. The network smooths most remaining ripple out of the dc. Also, the 200-ohm resistance, with current flowing through it, drops the output voltage for this branch to 190 volts dc.

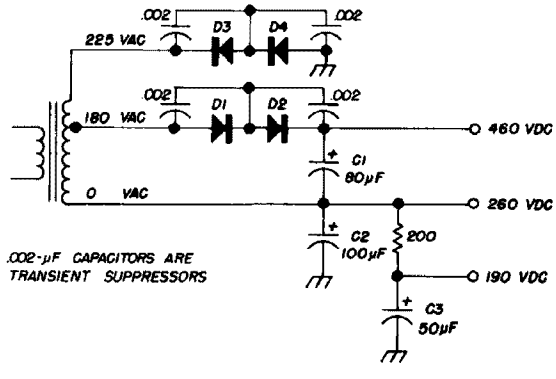
During the half-cycle when the top of the transformer winding is positive, the 180-volt center tap is also positive with respect to the bottom end. Rectifiers D1 and D2 can conduct with this polarity of voltage applied; they do, and charge C1. The peaks of these positive cycles reach about 250 volts. Capacitor C1 smooths out the pulsations and leaves an average dc voltage of about 200.

With respect to ground, C1 and C2 are in series, so the dc voltages across them add. The dc output is 260 volts (C2) plus 200 volts (C1), or 460 volts. The C1 voltage is developed during one half-cycle, and the C2 voltage is developed during the other—which makes the 460-volt output a result of full-wave rectification.

ssb on wheels

Single-sideband is a natural for mobile operation. Transmitting efficiency is high, which helps overcome the limitations of available power for the rig. In **fig. 10** is a supply that converts 12 volts dc to high-voltage ac and then back to even higher-voltage dc.

fig. 9. Interesting transceiver power supply provides three positive voltages.



The part of this circuit that turns dc into ac is called the inverter. The two pnp transistors are switches that alternate in conducting current, each through its own part of the primary winding. To begin the action, when 12 volts dc is first applied to point A, one transistor can conduct slightly better than the other, so it dominates. Suppose Q2 is that transistor.

Current flows through the winding from A to B to C and to ground through the 195-ohm resistor. A magnetic field builds up around those portions of the windings. The cyclic rise and fall of the magnetic field around the primary winding. That's ac.

The secondary of this transformer is a stepup winding for developing a high ac voltage. Its efficiency is improved tremendously by "tuning" it with a capacitor—the 470 pF. The output of this winding is about 1000 volts ac.

The rectifier diodes are series-connected in two banks of six each, called stacks. The series connection divides up the voltage across each rectifier. The two stacks make a full-wave doubler, producing an

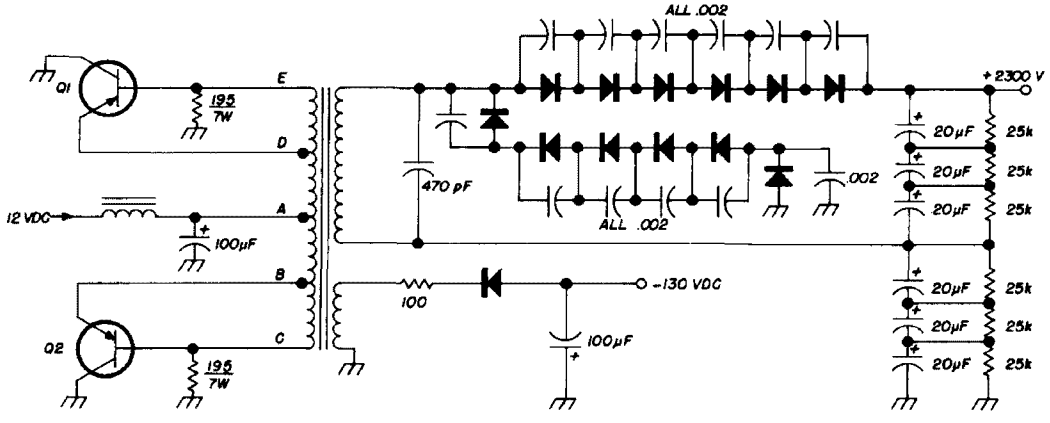


fig. 10. Mobile dc-to-dc converter that provides 2300 volts output from a 12-volt battery.

base of Q2 is a little less positive than the emitter, forward-biasing the pnp transistor. The collector, being grounded, is the most negative. Q2 conducts more and more heavily.

Whenever a magnetic field is expanding, it creates a counter-force that's opposing. When current in Q2 reaches saturation, the magnetic field stops expanding. The counter-force takes over and starts cutting down current in Q2 and in its winding sections. This reduction continues until Q2 is cut off by the reverse-bias that's self-induced by the winding.

Meanwhile, Q1 was kept cut off by the current rise that made Q2 conduct, because the two transistors are wired in opposite phase from each other. Once current in Q2 starts diminishing, however, current in Q1 starts rising. That makes current in the upper half of the winding increase, which raises the magnetic field around that half.

The result of this push-pull switching is a output of 2300 volts dc. As you probably guessed, the capacitors are wired in series—three to a stack—to permit using lower-voltage types. (High-voltage capacitors are very costly.)

The bottom secondary winding is connected to a negative-output rectifier, and supplies —130 volts ac. This is a bias voltage for the high-power ssb transmitter this power supply works with.

dc power for transistors

For operating transistors, input ac is first stepped down, then rectified. For critical transistor circuits, the low dc output may also be regulated. The power supply in **fig. 11** combines all three features.

There's something odd in **fig. 11:** transistors are used as rectifiers! This is the ac-input version of a supply that can also be used on 12 volts dc. The dc mode uses transistors Q1 and Q2 for inverting dc to ac. They're big, husky transistors, so why

not use them as rectifiers, too, and take advantage of diode action between their bases and collectors?

After full-wave rectification, the dc is smoothed by filter capacitors and a choke. The 12 volts dc is fed through R1 and R3 to the output terminal. The zener diode holds the base of Q3 at a steady voltage with respect to point A. Because of voltage drop across R1, current drawn through R3 makes the voltage vary at point A—and therefore at the base. The variation at the base makes shunt regulator transistor Q3 exactly the same amount—becoming that much more positive than normal. Voltage at the base becomes more positive, too, but not as much, because the voltage (and any change) is divided between R1 and R2. The net effect is to reduce the forward bias of Q2, lowering conduction. That reduces conduction in the base-emitter junction of Q1; that junction is in series with the collector of Q2. Less base current in Q1 raises its emitter-collector resistance. And that lowers the output voltage to its normal value.

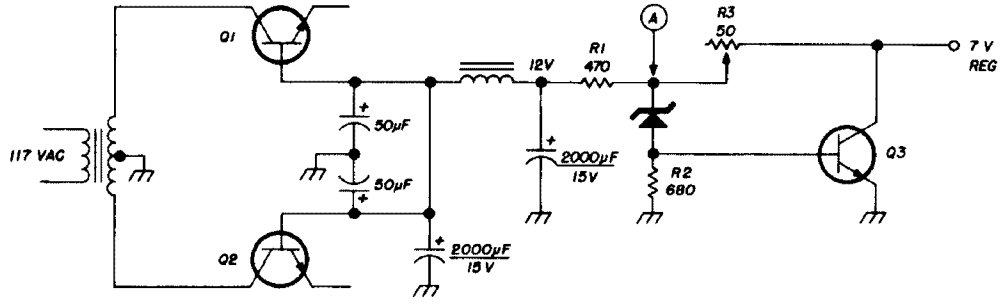


fig. 11. Regulated low-voltage power supply using transistors as rectifiers.

draw more current or less through R3 and thus hold the output voltage constant.

Consider what happens if the output voltage (and the voltage at A) rises. The base of Q3 goes more positive, and the transistor draws more current, reducing voltage at the collector (tied to the output). If the output voltage drops, the base goes less positive, and Q3 draws less current through the series resistors. R3 is adjustable to allow for aging of the zener or the transistor.

Transistors are used more often as series regulators. The example in **fig. 12** has an extra transistor, often known as an error amplifier. Operation is simple. Transistor Q1 is a series resistance, variable by the voltage at its base. This variable resistance parallels R2 in series with the output. The resistance of Q1 depends on conduction of Q2. An example illustrates how output voltage variations are corrected.

If the 12-volt output goes up, voltage at the emitter of Q2 goes up, through D2, by

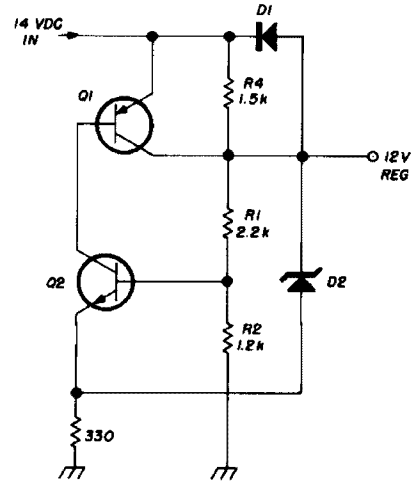
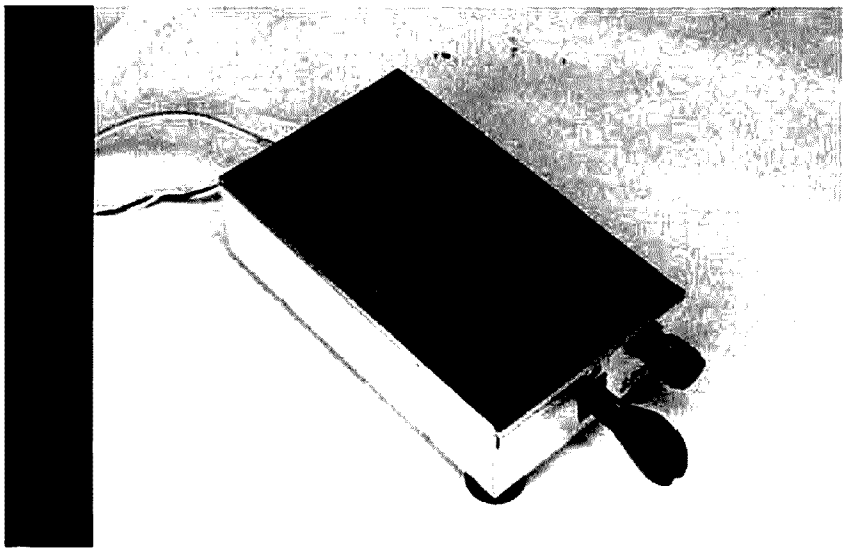


fig. 12. Solid-state series-type voltage regulator.

If the output goes down, forward bias is increased for Q2, and thus for Q1. The lowered emitter-collector resistance of Q1 lets the voltage go back up to normal. The result is a steady 12 volts, no matter what tries to lower it.

ham radio



Mini paddle installed in a homebrew keyer.

the mini paddle

Here's a simple low-cost keying paddle for the new electronic key you are putting together

Del Crowell, K6RIL, 1674 Morgan Street, Mountain View, California 94040

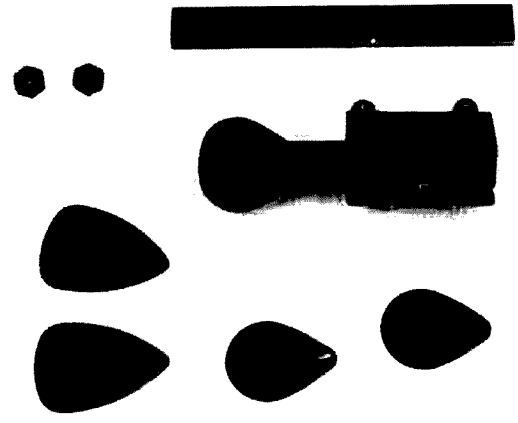
Many electronic-keyer circuits have appeared in the past several years, and they're getting so simple that almost anyone can build one. However, the paddles are a different story. The paddle described here requires a minimum of tools and is inexpensive to build.

construction

The paddle should be built with good quality materials. The arm is built from a section of 1/8-inch double-sided epoxy board; the sides must be flat and smooth. You can use aluminum or other material, but it will require much more work.

A pair of guitar picks are epoxied to the

Disassembled mini paddle.



paddle to provide a good surface for the fingers. By carefully using a drill, the bearing holes can be recessed in the edges of the epoxy material. For accurate alignment, these holes must line up with the holes in the bracket. The bearings can be salvaged from an old broadcast variable capacitor. Silver contacts removed from an old relay are soldered to the copper side of the paddle.

The bracket should be built from .040 brass, although tin or steel can also be used. side of the paddle. By adjusting the thickness of the spring you can adjust the tension to your own desires. The spring is soldered in place after adjusting. The paddle is held in place by bending the bracket to provide proper spring tension on the pivot bearings.

adjustment

Adjustment of the paddle is quite simple once the spring is adjusted for centering and soldered to the bracket. The contacts can be

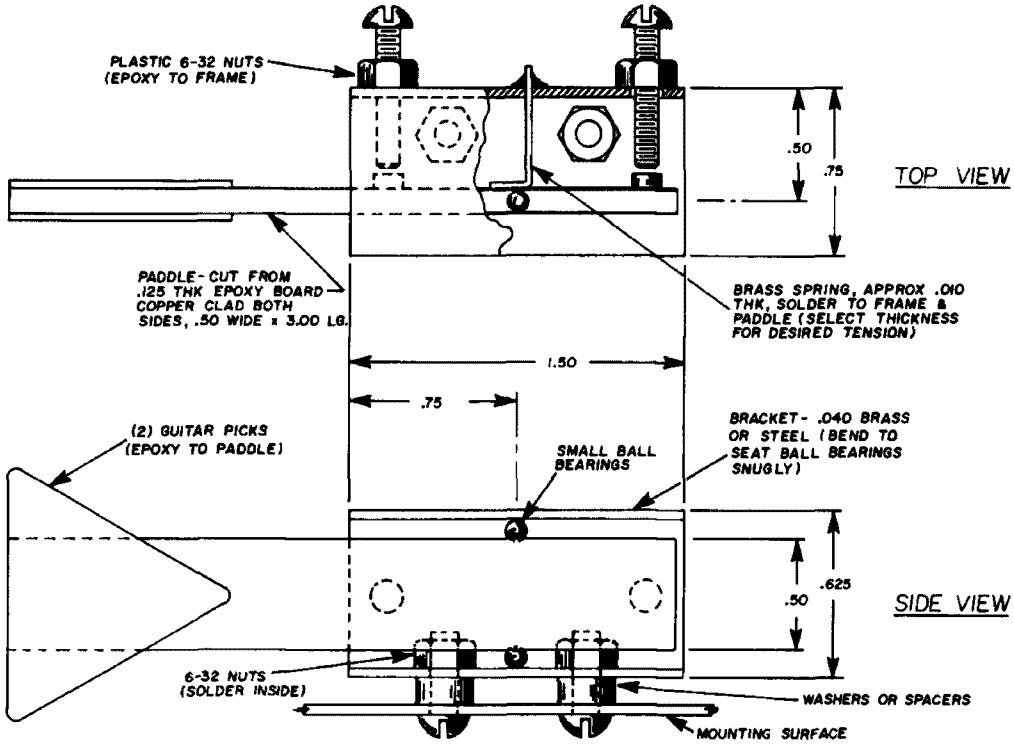


fig. 1. Constructional details of the mini paddle.

Mounting holes are located as shown in **fig. 1** and 6-32 nuts are soldered to the inside surface of the bracket. Guide nuts for the contact screws are provided by epoxying two plastic or nylon 6-32 nuts to the sides of the bracket. The clearance holes must be larger than the screw you use and the nuts must be centered to prevent shorting. Contact screws should be brass with the ends filed round and polished for good contact.

Spring tension is provided by installing a strip of brass material as shown in **fig. 1.** The size of strap is approximately 0.13 x 1 inch with one end bent and soldered to the flat adjusted to provide for proper travel and centering. I installed my paddle in an LMB 139 box along with the Micro TO Keyer described in the 1968 ARRL Handbook.

This keyer was originally built for mobile operation but it does a nice job as a second unit for the home station. It operates from a pair of flashlight batteries and has a built-in zener diode and dropping resistor so I can plug it into the cigarette lighter in the car. The monitor oscillator was eliminated for simplicity since my equipment has built-in sidetone.

ham radio

a sloping inverted-vee dipole

Here is an interesting antenna for small antenna farms that bears further investigation

Alf Wilson, W6NIF, 3928 Alameda Drive, San Diego, California 92103

Most of us at one time or another have dreamed of the ultimate antenna. If your ambition, for example, is to accumulate new country DX contacts, then this ultimate antenna might be a family of rhombics covering 360 derees, or maybe a 100-foot telephone pole with full-sized Yagis for each band. The majority of us never realize the ultimate antenna, unfortunately. As a matter of fact, there are hams who, for various reasons, will never be able to own and operate **any** kind of antenna but a dipole or one of its relatives.

It is for these hams that this piece is written. I hope they may benefit by my experience with a restricted-space antenna that evolved over several months of testing to achieve the best possible compromise. In general, those who are restricted to low-gain antennas fall into one or more of the following groups:

a. Thin pocketbook city dwellers on small lots

b. Not-so-thin pocketbook city dwellers on small lots who have zoning restrictions, unreasonable neighbors, unco-operative spouses or all three.

c. Timid city dwellers who are reluctant to live under a potential catastrophe such as a ton of tubular steel crashing onto the house

Many of these hams, knowing it is futile to participate seriously in the big contests, nevertheless would like to have a chance at working some DX at least once in awhile. Being antenna dreamers, these people have read extensively and have learned the pros and cons of horizontally- and vertically-polarized low-gain antennas. They have resigned themselves to using one of the dipole family; what, then, is the best possible arrangement they can build that will provide at least a fair chance in the DX bands?

Let's assume a typical city dweller has room to erect an inconspicuous mast at least one-half wavelength high on 14 MHz, but doesn't have enough room laterally to erect a second mast one-half wavelength from the first. Full-sized horizontal dipoles are therefore out. He could put up a shortened (trap) dipole or one of the verticals, say a ground plane. Both have advantages and drawbacks, all of which are well known and will not be discussed here. Which is most effective, considering the restrictions involved?

Based on these restrictions the antenna to be described, while not in the competition class by any means, is a modified version of the dipole family that has given a good account of itself over a six-month period on 14 MHz. It is a variation of the well-known inverted vee. The conventional inverted vee has been covered fairly extensively in the literature[1,2,3,4]. Very little has been published, however, on any but the standard configuration.

the inverted vee

The standard inverted vee is a half-wavelength dipole with the elements in a vertical plane, oriented in an acute angle, α, and fed at the apex (**fig. 1**). The size of the apex angle apparently has little effect on the vertical radiation pattern but significantly changes the horizontal (earth plane) pattern such that the nulls off the ends normal to the wire axis tend to fill in as the apex angle decreases. (Reference 5 discusses this at length and provides typical patterns in both planes for several apex angles.) Thus as the antenna approaches a vertical configuration, its horizontal pattern approaches the omnidirectional pattern of the conventional vertical, which is to be expected. Reference 5 shows that the vertical pattern is virtually unaffected for antenna apex angles from 120 to 60 degrees. Reference 6 shows the vertical radiation pattern of any antenna should be lower than 30 degrees for effective DX work.

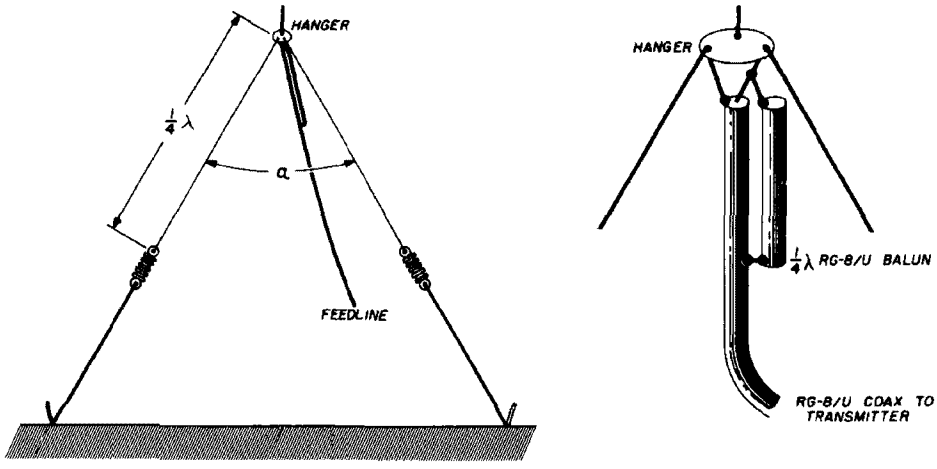


fig. 1. Basic inverted-vea antenna. Resonant frequency depends upon height above ground and angle. The length of the quarter-wave balun is equal to 246v/f, where v = 0.659 if balun is taped to leadline, and f is frequency in MHz. Center conductor of the "bazooka" section is floating.

The problem is to somehow get the vertical pattern down into the useful angles without sacrificing energy in high-angle lobes that penetrate the ionosphere. Increasing the antenna height above one-half wavelength does lower the vertical angle but introduces high-angle lobes that are useless for DX propagation. Also, recall that we are restricted to one-half wavelength in height on the 14 MHz band.

the sloping inverted vee

After using a conventional inverted vee at a height of one-half wavelength on 14 MHz for a month or so, I wanted to find out what would happen to antenna performance if the elements were rotated out of the x, y plane. The elements were kept broadside to the x, y plane but were elevated at a slope angle, θ (**fig. 2**).

At this point I must emphasize that no amount of wishful thinking will alter a basic physical fact; namely that the directional characteristics of a true vee antenna cannot be realized with anything less than one wavelength on the legs. A vee-shaped dipole is exactly that: its shape is that of a vee, but there its similarity to a vee antenna ends. The inverted vee dipole behaves like a dipole regardless of the apex angle, α, or the slope angle, θ. However, it appears that if the slope angle is between 40 and 70 degrees, the vertical radiation pattern angle of the lobe in the direction opposite to that in which the vee is rotated tends to decrease, so that more energy is radiated at angles below 30 degrees.

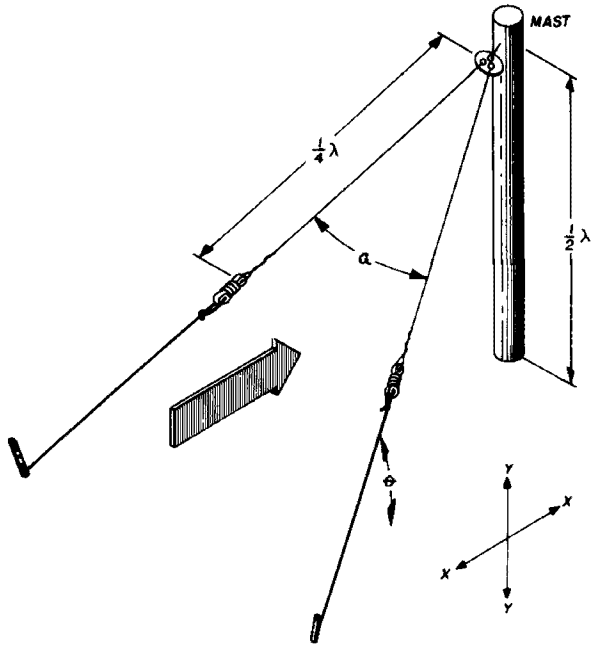


fig. 2. Geometry of the sloping inverted-vee antenna. The angle of the vertical lobe decrease in the direction of the arrow.

table 1. Characteristics of several simple antennas; all assumed one-half wavelength high and fed with the same current.

antenna	nominal radiation resistance (ohms)	vertical pattern useful lobe angles (degrees)	first reflection zone* (miles)
horizontal dipole	73	15 – 45	200 – 1100
ground plane	53**	10 – 55	100 – 1500
inverted-vee dipole	50	15 – 45	200 – 1100
sloping inverted-vee dipole	50	15 – 25 (estimated)	600 – 1500 (estimated)
yagi	8	15 – 45	200 – 100

*for F-layer heights between 125 and 250 miles
**with radials sloped approximately 50 degrees

tests

My sloping inverted vee dipole was oriented with the elements broadside to the long-path direction (about 230 degrees azimuth for my location). Over a six-month period, the antenna was used in tests with South African, Australian, Southeast Asian, and Asian stations on 14 MHz. Approximately 100 contacts were made. During these tests the sloping inverted vee antenna was compared with a shortened dipole and a quarter-wave ground plane. The driving points of all three antennas were within about 5 feet of being at the same height above ground. The inverted vee was rotated out of the x, y plane to slope angles as great as 70 degrees. This was accomplished by rigging a lightweight spreader between the antenna elements and raising

them to various angles with a rope and pulley arrangement. An interesting effect was that the optimum slope angle seemed to depend on the time of day. During band openings on 14 MHz (1300-1400 GMT) best results were obtained with a slope angle of about 40 degrees. As the sun rose, reports improved with the slope angle approaching 70 degrees.

results

No quantitative measurements were made. Pattern and field strength measurements are difficult enough even under laboratory conditions; consequently any data taken with the limited facilities available would certainly be open to question. However, after testing the antenna under all band conditions, the following observations are offered:

a. The sloping inverted vee antenna apparently radiates better in the vertical plane at lower angles than a dipole or ground plane.

b. The first reflection zone of the sloping inverted vee appears to be at a greater distance than the comparison antennas.

c. As a receiving antenna, the sloping inverted vee is noisier than the dipole, but not as noisy as the ground plane.

d. Band openings occur earlier by about a half hour with the sloping inverted vee.

e. The sloping inverted vee requires less space than the dipole, but more than the ground plane.

Based on the data in reference 6 and the empirical results of the tests described above, the characteristics of the sloping inverted vee are summarized in **table 1**.

conclusions

In conclusion I would like to point out that the sloping inverted vee is not in the same league as even a two-element Yagi, which has both directional and power gain. While the sloping inverted vee does seem to have more power at the lower angles than a conventional horizontal dipole, it simply cannot compete with a Yagi or any other directional array. The estimated first reflection zone for the sloping inverted vee is based on the curves of reference 6, which in turn are based on various vertical pattern lobe angles. Experience with this antenna seems to indicate performance close to the data shown in the table.

I hope that others will try this antenna with different slope angles. It will be very interesting to see how it performs at other locations and on different bands.

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ham radio

vtvm modification

Most garden-variety vtvm's use a 1.5-volt flashlight cell for the ohmmeter supply. This works fine if you replace the flashlight cell every few months. However, the ohmmeter becomes very inaccurate on the low-ohms scale as the cell's internal resistance increases with age.

By adding a few parts and making a few circuit modifications, it is possible to do away with the flashlight cell altogether. The diagrams show the modifications I made to my Heathkit V-7A. This is typical of all inexpensive dc vtvm's, whether made by Heath, Allied, Eico or RCA.

As shown in **fig. 1,** the 6.3-volt filament voltage is rectified by a conventional bridge rectifier; the resulting dc is dropped by the 18-ohm resistor in conjunction with the regu-

lating diodes, giving the required 1.5 volts dc.

It's necessary to lift one side of the heaters from ground before you can use the 6.3-volt winding in a bridge rectifier. Heathkit uses a printed-circuit board, but the conductors that must be removed can be easily scraped off the board with a sharp knife.

Two 1N538's in series act as a zener to regulate the output voltage. The barrier potential for a silicon diode is about 0.75 volts; two in series regulate the ouput nicely at 1.5 volts.

The original flashlight cell was connected to a resistor string tapped by the ohmmeter range switch as shown in **fig. 2.** Notice that the total resistance switched into the circuit is always 10 times the range switch position; this assumes cell internal resistance of 0.9 ohms. The internal resistance of this power supply is higher than 0.9 ohms, so it's necessary to lower the 9.1-ohm resistance by shunting it with 35 ohms as shown. The two in parallel then make 6.75 ohms. When this is added to the power supply internal resistance of 3.25 ohms, it makes up the required 10.0 ohms.

In my vtvm I also replaced the 6AL5 "ac volts" rectifier with a pair of high back resistance, 600-PIV silicon diodes. This saves about as much heater current as consumed by the 1.5-volt power supply, so the transformer is still running cool.

The diodes that replace the 6AL5 should be forward biased slightly to overcome their barrier potential. This is done with a voltage divider consisting of a 24k- and 150-ohm resistor connected to the negative side of the original plate supply (**fig. 3**). Linearity is rather poor below levels of 1 volt ac, but is still superior to the original 6AL5.

Fred Brown, W6HPH

fig. 1. Deriving a 1.5-volt dc supply for the vtvm from the 6.3 Vac filament transformer. Added circuitry is enclosed by the dotted line.

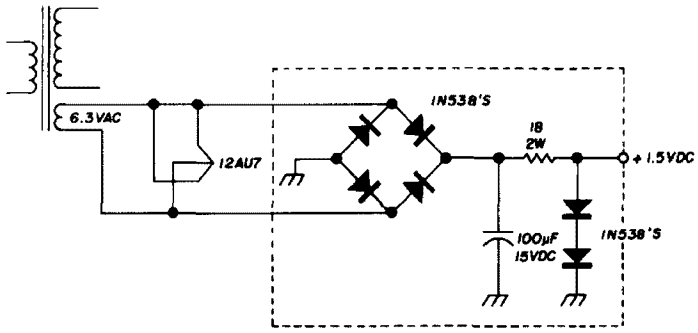


fig. 2. Modifying the ohmmeter circuitry for the 1.5 Vdc supply.

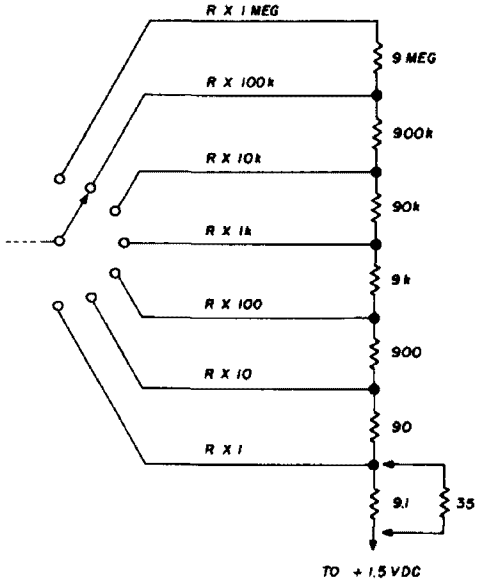
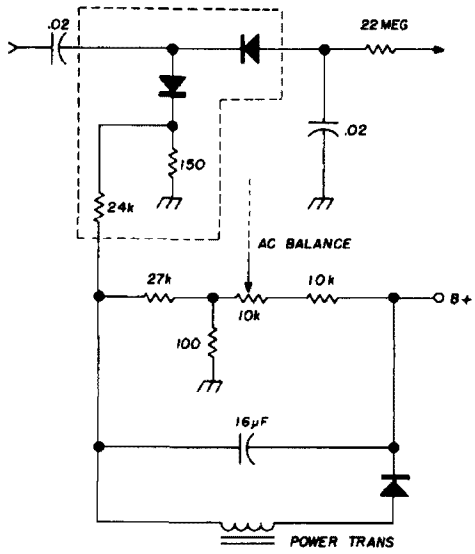
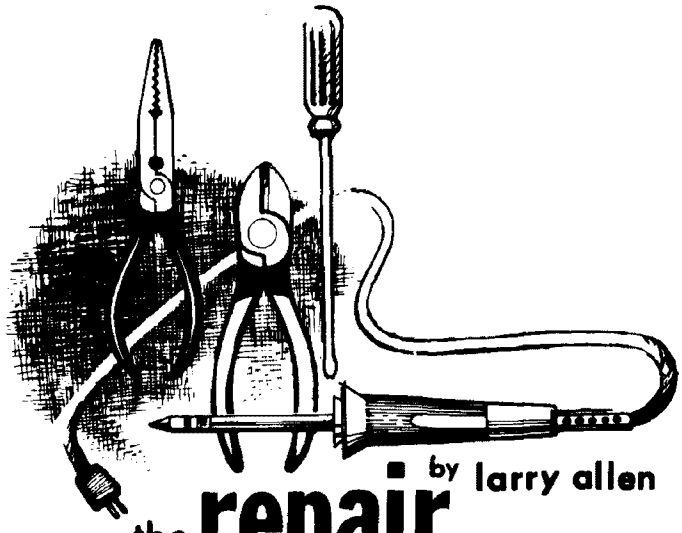


fig. 3. How to replace the 6AL5 with forward-biased silicon diodes.





the repair bench

by larry allen

analyzing wrong dc voltages

There isn't really a "best" way to troubleshoot electronic equipment. The most successful radio and electronic technicians combine several methods. I've already told you about some of them in past columns, but this month's method is probably the most popular; at least it's one you have to understand if you expect to repair every trouble in all sets. It's called **voltage troubleshooting.**

Just because this method is the best known doesn't mean it's always best. Most experienced guys save this one for last. It's a quick way to find the faulty part in most circuits, once you know which circuit to look in.

You can find which circuit this way too, but other ways are usually quicker. I'll use this month's column to tell you how to pin down faulty parts within the circuits; that's the best way to use voltage troubleshooting.

how the method works

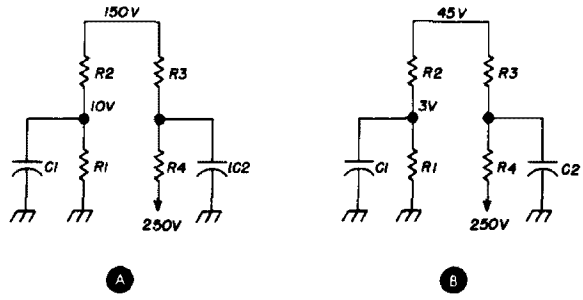
In theory, it's simple. You start by measuring each dc voltage in the suspected circuit. Then compare yours with correct voltages. Most instruction books and schematics nowadays include voltages—at least at each tube or transistor terminal. When your circuit diagram doesn't show the *working voltages, use past experience and* your knowledge of how the circuit works.

When you find a voltage that isn't right, figure out what could cause it. If you know Ohm's law for voltage, current, and resistance, it's not too hard to decide what's causing the voltage to change. In fact, it's really cut-and-dried.

Here's an example. In **fig. 1** are two versions of a simple series resistance circuit connected across a dc voltage. The first, **fig. 1A,** shows normal dc voltages (which I'll call operating voltages). The other, **fig. 1B,** shows voltages you might measure in this circuit when you start troubleshooting. (Voltages are always measured with respect to ground.)

There's 150 volts at the R2-R3 junction in **fig. 1A** because current flows through R3 and R4, causing (says Ohm's law) a drop of 100 volts. There's only 10 volts at the

fig. 1. Simple series resistance circuits show how voltages divide according to the various resistance ratios.



R1-R2 junction because another 140 volts is dropped across R2. The same amount of current flows in all four resistances, since they're in series, so the voltages divide up exactly according to the resistance ratios.

You analyze voltage to determine why the two voltages in **fig. 1B** have changed from normal. What part has gone bad? For some reason, the resistance ratios must have changed. Your first step, then, is to figure out what the new ratios are.

If you consider the 45 volts, you can quickly see two possibilities. The first is that R1 or R2 may have lowered in resistance. In that case, the ratio of their resistance to that of R3-R4 lowers, and less of the total voltage appears across them. On the other hand, the combined resistance of R3-R4 may have increased. That causes more voltage to drop across them and lowers the voltage reaching the top of R2.

So . . . which is it? The clue lies in the fact that **both** the R1-R2 voltage **and** the R2-R3 voltage have kept the same ratio between them, even though both are lower than normal. The 3 is to 45 as the 10 is to 150; the ratio is still 1 to 15. The trouble is more likely in R3 or R4; the value of one of them must have increased. To find out which one, you could disconnect them and measure with your ohmmeter. However, further voltage measurements will show one of them still retains its ratio to R1-R2; the other is the culprit.

This simple reasoning is the basis for analyzing all dc voltages in series circuits. You study the ratio of resistances, compare the ratio of voltages, and then figure out what's causing the foulup.

voltage dividers and ohm's law

You should already know about series and parallel circuits.* For voltage troubleshooting, you must know how voltages spread themselves around both kinds of circuits.

Start with the simple series circuit in **fig. 2A.** The resistors divide the voltage in proportion to their values. The ratio is 2:2:1. The R1-R2 junction is 80 volts below the supply voltage, and the R2-R3 junction is 80 volts further below supply voltage. Looking at the divider string from the standpoint of the voltmeter: R1 drops 80 volts, R2 drops 80 volts, and R3 drops 40 volts. The 2:2:1 ratio is maintained.

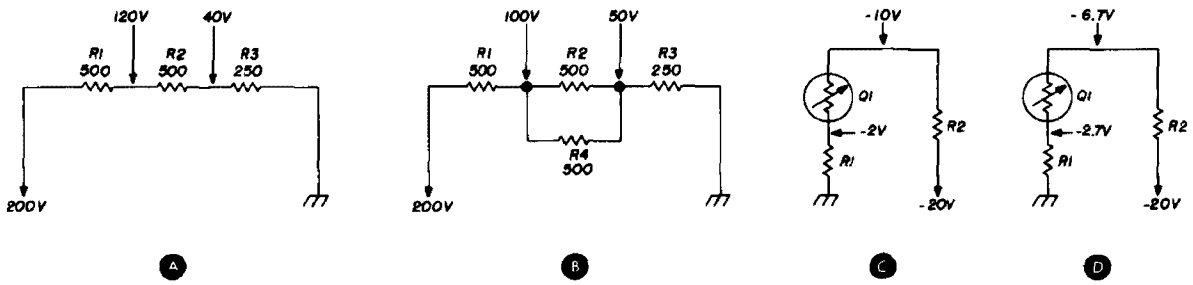


fig. 2. How voltages change when one of the resistance values changes.

Look what happens when you alter a resistance value, as in **fig. 2B.** Some additional resistance, in the form of R4, has been added in parallel with R2. It lowers the effective resistance; R2 and R4 are both 500 ohms, so their parallel value is 250 ohms.

That goofs up the resistance ratio of the divider chain. It has become 2:1:1, and the voltage divides accordingly. R1 drops 100 volts, R2 (with R4) drops 50 volts, and R3 drops 50 volts. So you see, among the voltages, the 2:1:1 ratio holds.

Now, imagine that **fig. 2C** is a supply circuit in a solid-state receiver. The variable resistance labeled Q1 is actually the tran-

* For a rundown on series and parallel circuits, check the **repair bench** in the November, 1968 and January, 1969 issues of **ham radio.**

sistor, but imagine it here as a variable part of the divider chain. Voltages are shown at each point in the circuit, so you can figure out the voltage (and therefore resistance) ratio. Across R1 is dropped 2 volts, across Q1 is dropped 8 volts, and across R2 is dropped 10 volts. The ratio is 1:4:5.

Suppose something goes wrong with the circuit. You measure with your voltmeter and find the voltages changed to those in **fig. 2D.** Now you have a new ratio of voltages. Across R1 is 2.7 volts, across Q1 is 4 volts, and across R2 is 13.3 volts. If you write out the new ratio, however, it's easy to see where the voltage changed. The new ratio is 2:3:10. The old ratio was 1:4:5, which is the same as 2:8:10; only the center factor has changed. That means the resistance of Q1 has changed, and that's what upset the voltages.

tubes and transistors as dividers

Thinking in ratios is easy as long as the resistances are simple. Experience will teach you to estimate voltage ratios close enough to give you a clue to which resistance has changed.

Tube and transistor circuits are different. Changing the bias on either one alters its dc operation. **Fig. 3** suggests how to consider a tube or transistor as a variable resistance in the dc supply circuit. In **fig. 3A,** the tube bias can be changed by moving the slider on cathode-bias resistor R1. The equivalent circuit shows the tube's dc plate resistance as variable; it actually changes as bias is varied.

Simplicity is lost because R1 is also in series with the dc supply circuit. When the resistance of V1 changes, current changes all through the series circuit. A change in current through R1 affects bias, which changes the plate resistance of V1, which changes the bias, which changes plate resistance, and so on. The interdependency makes it tough to treat tube circuits as simple ratio dividers.

This problem also applies to transistor circuits; to see how, check **fig. 3B.** Bias for the transistor base is controlled by voltage divider R1-R2; since R1 is a potentiometer, bias is variable. As bias is altered, more or less current can flow in the emitter-collector circuit. The effect is to alter the emitter-collector resistance, represented by resistance Q1 in the equivalent circuit.

To see how complex these relationships can get, take a look at **fig. 4**—which is from a ham rig. Equivalent diagrams are shown at the right. The one at **fig. 4B** is for the collector circuit; it's the supply path through which dc must flow. **Fig. 4C** is the dc supply path for the base circuit; it in-

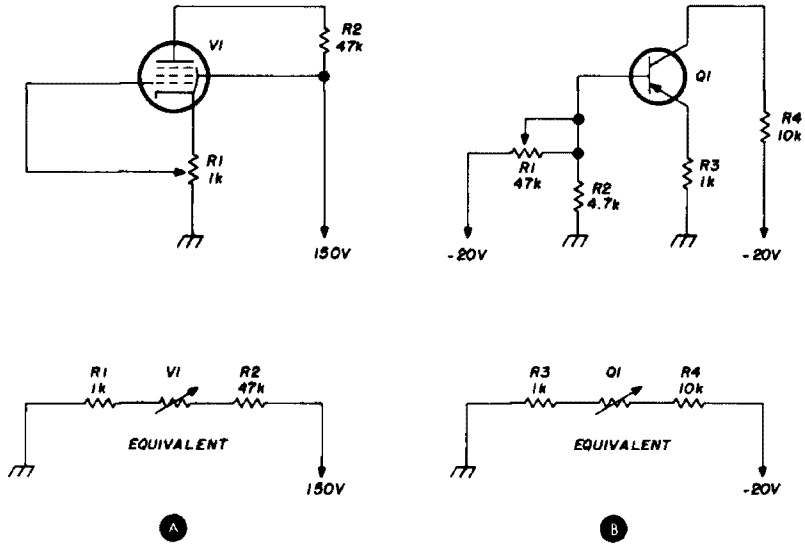


fig. 3. Tube and transistor circuits with their equivalent resistive divider circuits. Drawing the equivalent circuit makes it easier to see the effect when a resistance changes.

cludes the biasing voltage divider and the path through the base-emitter junction (shown as Q1B). Notice that R3 is in this path, too, so a current change in either circuit changes voltage across that resistor.

Finally, notice the combined paths in **fig. 4D.** Figuring out ratios in this circuit would be hard even if the resistances were simple. But they're complicated by the fact that the value of Q1C is controlled by Q1B. It's just too much to be simple.

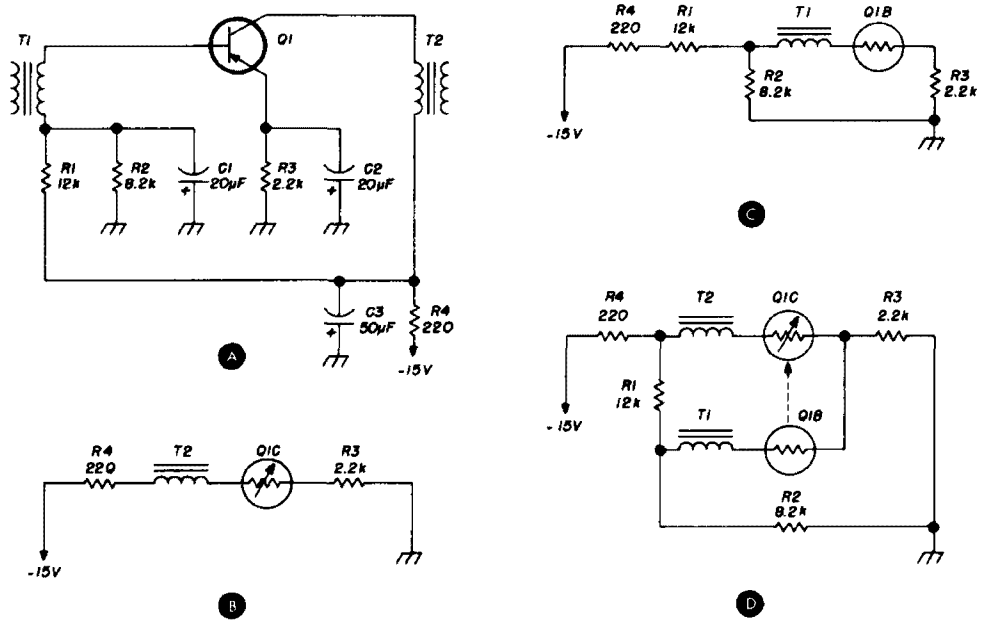


fig. 4. Transistor audio amplifier; the equivalent circuits in B, C and D show how difficult it is to use the "ratio analysis" mathod of troubleshooting a stage like this.

a step at a time

However—and this is important—even though you can't use voltage ratios directly in tube and transistor circuits, you can apply the principles. Concentrate on one supply path at a time. At each step, ignore the effects of other circuits. When you decide what happened in one circuit, then study whether another could possibly be causing the voltage upset.

The transmitter tube stage in **fig. 5** is an easy example. Suppose you've traced trouble to this stage and have decided to measure voltages to help you find the faulty part. Your voltmeter tells you the plate voltage is low—only 180 volts instead of the 235 called for by the schematic. All the other voltages are about normal.

The plate-supply circuit (all the resistances plate current flows through) includes R5, one winding of T1, the tube itself, and R3. One at a time, consider the effect each part **might** have on voltage at the plate. Redrawing the equivalent plate-supply circuit, as in **fig. 5B,** can help. Here's the way your thinking might go.

If R3 increases in value, the dc voltage at the plate increases instead of decreases. If R3 decreases, the voltage lowers—but only a little, since R3 is such a small portion of the over-all circuit. Conclusion: forget about R3 for the moment.

If the V1 resistance increases for some reason, the dc plate voltage goes up. However, if the resistance of V1 decreases, the plate voltage goes down—exactly the symptom you have. Conclusion: the trouble might be V1. But don't rush; check other possibilities, too.

The winding of T1 can be eliminated as a suspect. If it opens, there's no voltage at the plate of V1. If it shorts, it changes the

voltage too little to notice. If the value of R5 increases, the voltage at the plate of V1 lowers. So, there's another suspect! If R5 decreases in value, voltage goes up instead of down.

Now you have two suspects: R5 may be too high in value, and the dc plate resistance of V1 may be too low. Which is it? To find out, you measure other voltages in the stage.

For example, what about screen voltage? It's okay. If the dc plate resistance of V1 is low, it also lowers the voltage at the screen. But it isn't, so the tube and its bias must be okay. (You can make sure bias is okay by measuring it.) That leaves R5 as the likely culprit. You can just measure R5 with your ohmmeter to be sure.

Suppose you do find the screen voltage low, too. What then? The cause is probably the tube plate resistance, but what's making it low? It could be a faulty tube or incorrect bias. Either one could allow too much current through the tube, and thus lower the plate resistance. So—measure the bias. If it's okay, the tube must be bad. If the bias is wrong, go into that circuit to find out why.

The secret to voltage troubleshooting is to take it step-by-step, through each portion of each dc supply circuit. Decide how each part could cause the symptom you observe. If you trace the trouble to the tube or transistor, find out if it's bad or if some other operating voltage is making it work wrong.

voltages in transistor stages

The emitter-collector resistance of a transistor is set by bias current in its base circuit. Just as plate voltage of a tube may change if bias isn't right, the collector voltage of a transistor can, too.

The transistor stage in **fig. 6** is a mixer from a communications receiver. The transistor is a pnp type, forward-biased (base more negative than emitter). Dc operating voltages are listed alongside the emitter, base, and collector leads.

Suppose you trace trouble to this stage and want to use voltages to pin down the faulty part. The voltages you measure are listed in **fig. 6B.** Collector voltage is high, suggesting there's either too little current through the transistor or R4 is too low in value. If the trouble is in R4, however, the voltage at the emitter must also be high —and it isn't.

The best clue is the positive voltage on the base. It is reverse-biasing the base-emitter junction, cutting off collector current through the transistor. You have to trace the cause of the positive voltage. As it turns out, the other side of C1 goes to a positive voltage in the preceding stage; C1 is leaky, and is coupling the wrong voltage to the base of this stage.

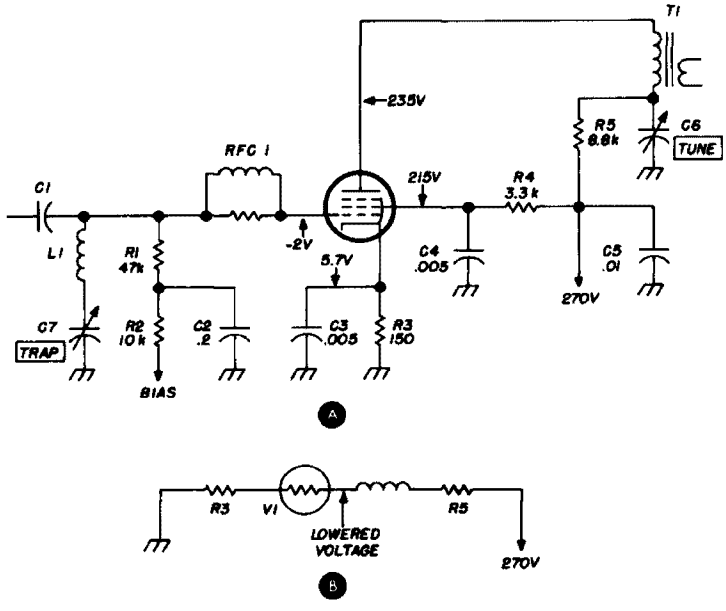


fig. 5. Driver circuit, with normal voltages listed at tube pins; B show equivalent plate circuit for troubleshooting.

Suppose, instead, you measure the voltages listed in **fig. 6C.** Collector voltage is very low, meaning there's excess current through the transistor or R4 is too high in value. Base voltage is a little low, but it hasn't changed nearly as much as emitter voltage has. Study the circuit. You'll quickly decide that C2 has shorted. With 1.9 volts of forward bias (instead of the normal 0.5), the transistor conducts heavily, lowering collector voltage drastically. The base voltage is slightly lower because of

the effect of heavy current through R4; the base also gets its voltage from R4, through R2.

There are countless transistor-supply circuits you can analyze this way. Always pick out the voltage that is most wrong and find the cause of that first. Then check possible causes of other wrong voltages. The part that's common to all symptoms is usually the culprit. Each symptom leads to another, and ultimately to the bad part.

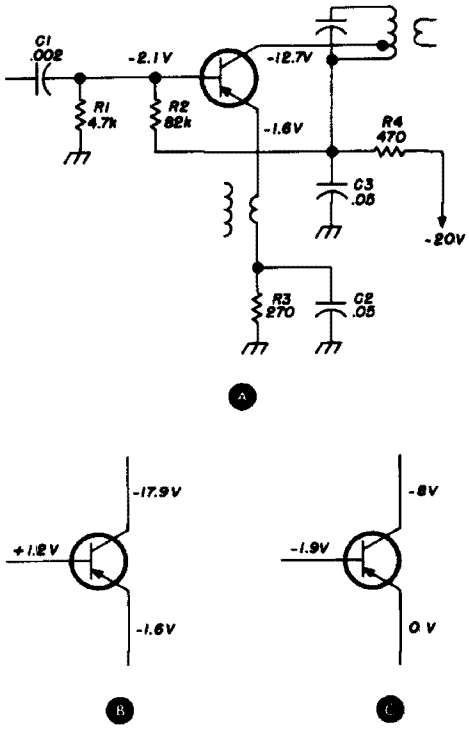


fig. 6. Transistor receiver mixer circuit; B and C show voltage changes in different troubleshooting situations.

testing individual components

This is the end of the line in troubleshooting—testing individual parts to make sure they're really bad. Resistors you can test most easily with your ohmmeter. Capacitors and other parts can be evaluated by voltage tests, combined with simple logic.

For example, C1 in **fig. 6** should couple absolutely no dc voltage from the preceding stage. If the preceding voltage is positive and the capacitor is leaky, the result is what you get in **fig. 6B.** If the preceding voltage is negative, it could drive the base of the transistor more negative and wind up reducing collector voltage. Leakage through C1 might also reduce voltage on the preceding stage—another helpful clue.

There is a reasonably sure way of testing a capacitor for leakage. If one end is connected to a dc voltage, disconnect the other end. With a vtvm, measure there for dc voltage. If voltage is leaking through, the voltmeter will detect it.

Don't be misled by electrolytic capacitors. They have a certain amount of natural leakage. Best way to check them is with an ohmmeter; leakage resistance should be no lower than 30k–50k.

You can tell if a coil or transformer winding is open, provided the winding carries dc. Just check at both ends with your voltmeter. An open winding blocks voltage.

You can also check for leakage between windings. Make sure one winding is connected to a voltage. With the other winding completely disconnected from any circuit, touch the voltmeter probe to one of its wires. If you measure any dc voltage, the transformer is leaky.

Any short-circuit can be pinned down quickly. Just find where the voltage has dropped completely to zero. This technique is most helpful in supply circuits where a filter, a bypass, or a high-power tube has shorted.

coming up

Awhile back, I promised to devote a column to using an oscilloscope at the repair bench. Apparently more amateurs own scopes than I suspected, but not many know how to use them or for what. Of course, everyone (just about) knows how to hook a scope up as a modulation monitor for a-m. But there are dozens of uses for this versatile instrument.

In my next column, I'll tell you exactly what a scope is and how to go about using it most easily. I'll give as many uses for it as I have room for. Then, after that, I'll explain a bunch of troubleshooting shortcuts with the scope. If you've got one, you should be using it on the repair bench. It's doggone handy.

ham radio

propagation predictions for february

Special tabular predictions for the ARRL DX contest

Victor R. Frank, WB6KAP, 12450 Skyline Boulevard, Woodside, California 94062

In the past, I have stayed clear of forecasting propagation for specific paths. Forecasts have been as general as possible so they would be applicable to any of the paths that might interest a radio amateur in the 48 contiguous United States. The format I have used makes you "work," but you will have obtained more accurate forecasts than otherwise possible.*

Still, the idea of doing a special propagation forecast for the CQ World-Wide DX contest or ARRL DX contest with the opening times tabulated for many paths **was** appealing. Now this copy is way past deadline, a $46 computer bill has just arrived and I am swearing off! No more worldwide specific path predictions—at least until the next contest.

If you live within 300 miles of San Francisco, Dallas or New York City, and are interested in knowing when ten meters will open to Outer Mongolia or 18 other rare or not-so-rare DX spots, the tables are made for you. If you live between San Francisco and Dallas or between Dallas and New York you can interpolate the path opening times on many paths (not over the pole). Otherwise, the usual propagation charts, **fig. 1** through **6** will suffice.

* If you're looking for propagation forecasts for specific paths, there's always George Jacob's propagation column in CQ magazine.

Upon investigating the tabular path predictions, you will find, perhaps to your amusement, that opening times have been predicted to the nearest third or half hour on the lower bands and to the nearest 10 minutes on the higher bands.

Lest I be called to task for missing an opening by 10 minutes, or even a half hour, let me point out that these are median short-path predictions, and that variations of 10 percent in path muf and 10 dB in absorption are not uncommon. In addition, station parameters may vary considerably from the 100 watts CW or 800 watts ssb assumed on 80 through 10 meters (12 watts CW or 100 watts ssb on 160 meters). Antenna gains (over an isotropic radiator) are assumed to be −6 dB for 160 and 80 meters, zero dB for 40 meters and +6 dB for 20 meters and higher frequencies.

The ten-meter paths with no openings forecast **may** be open during the middle of the fifteen-meter opening—because the muf may be higher than forecast or the long path may be "open." There are some marginal openings forecast that may catch the average contester unaware, such as the reopening of 20 meters after midnight or the short over-the-pole openings near dawn and dusk.

The serious contest operator will be prepared with at least one extra receiver, perhaps a panadapter, as well as a receiving antenna for checking "dead" bands. However, a serious weekend of bandscanning and **listening** just before the big contest can give more useful propagation information than any column.

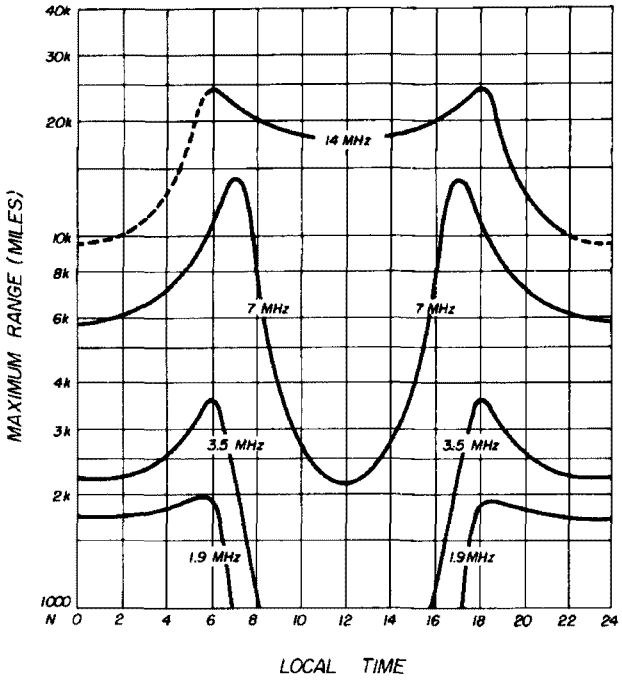


fig. 2. Maximum range to the north from 38° N latitude due to absorption.

six meters

Judging by last fall's openings, six meters is expected to open infrequently (about 10

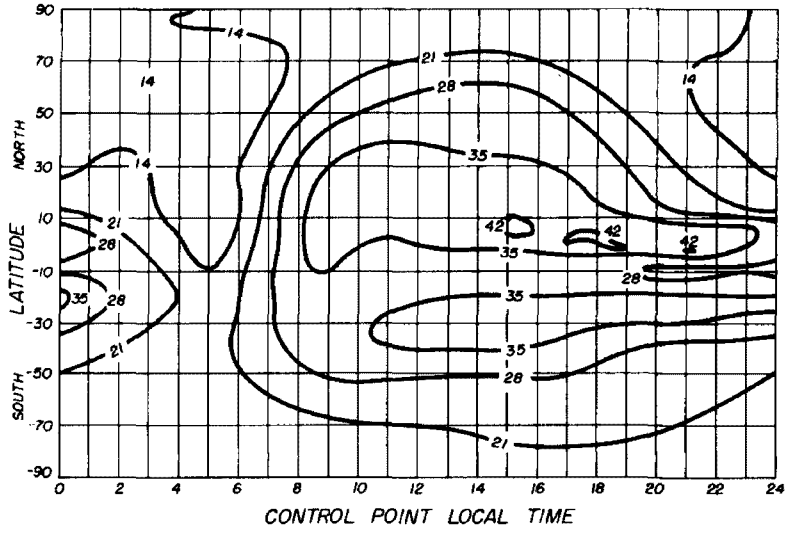


fig. 1. Maximum usable frequency curves for February 1969 based on 90° W longitude.

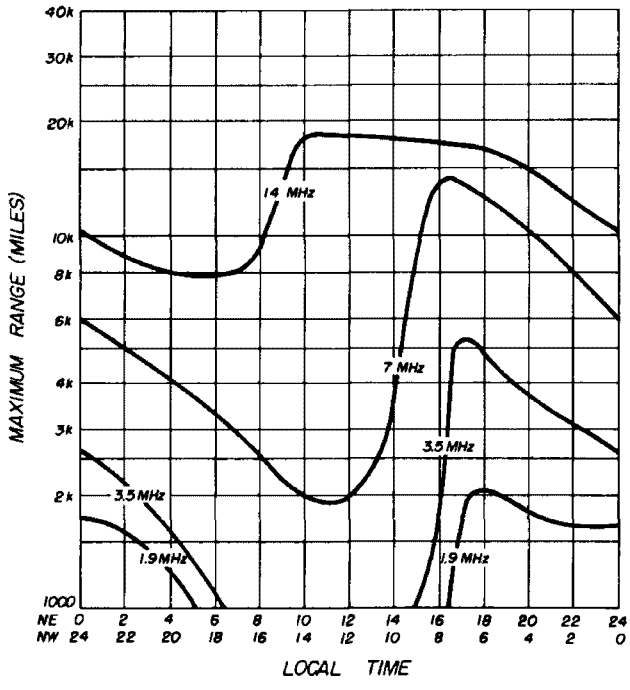


fig. 3. Maximum range to the northeast (top time scale) and to the northwest (bottom time scale) from 38° N latitude due to absorption.

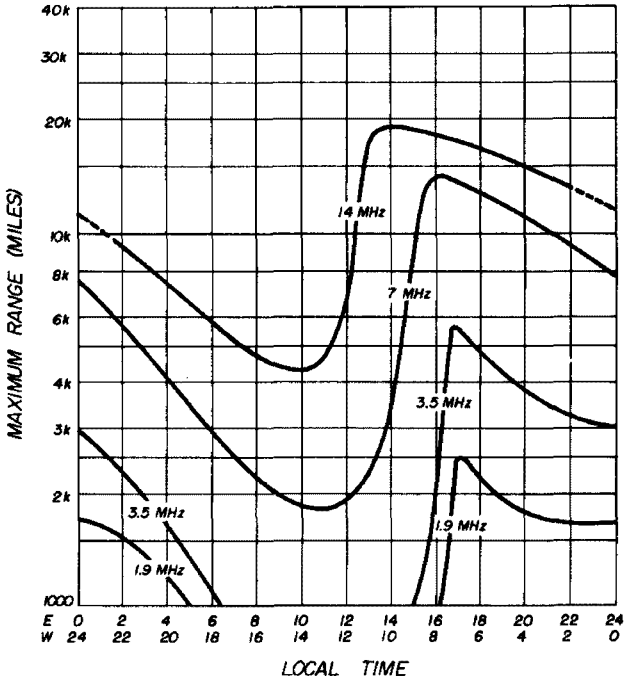


fig. 4. Maximum range to the east (top time scale) and to the west (bottom time scale) from 38° N latitude due to absorption.

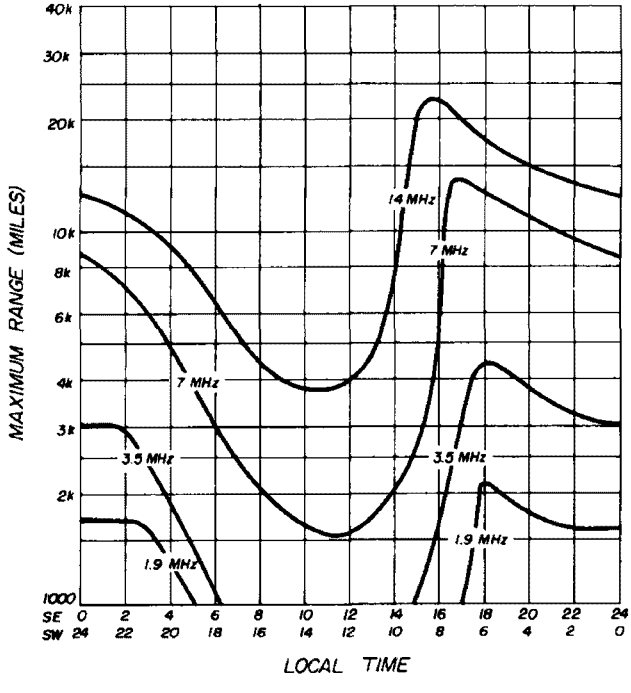


fig. 5. Maximum range to the southeast (top time scale) and to the southwest (lower time scale) from 38° N latitude due to absorption.

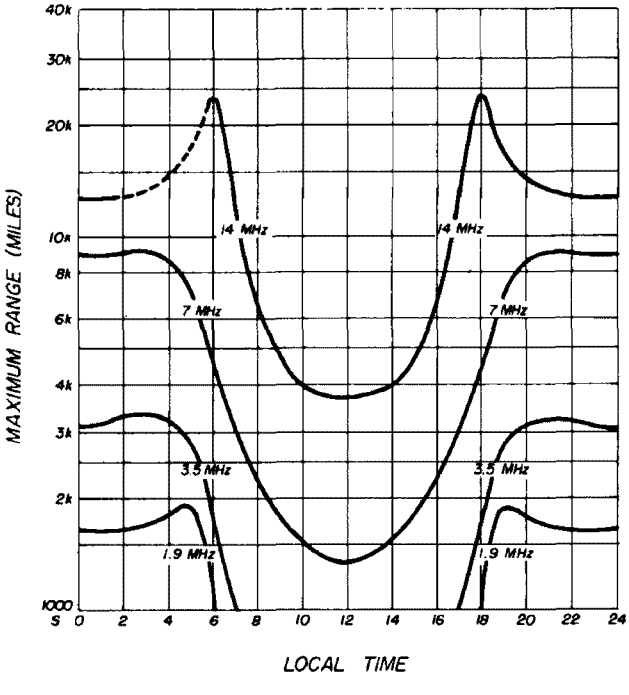


fig. 6. Maximum range to the south from 38° N latitude due to absorption.

table 1. Predicted 160-meter openings during the ARRL DX contest (time in GMT).

to from	San Francisco	Dallas	New York
XE, Mexico City	0200-1200	0030-1240	0100-1100
KP4, Puerto Rico	0200-1000*	0100-1000	2300-1000
KH6, Oahu, Hawaii	0500-1400	0500-1300*	0500-1100*
KL7, Anchorage	0400-1400	0300-1300*	0300-1100*
KZ5, Canal Zone	0200-1100*	0100-1100	0000-1000
OX, Thule, Greeland	0200-1230*	0100-1130*	2300-1000
PY, Belem, Brazil	0200-0900*	0100-0900*	2300-0900*
ZK1, Cook Islands	0600-1400*	0600-1200*	0900-1100*

* Difficult path, 10 dB additional circuit gain assumed.

table 2. Predicted 80-meter openings during the DX contest.

to from	San Francisco	Dallas	New York
XE, Mexico City	0120-1320	0000-1330	0000-1200
KP4, Puerto Rico	0100-1100	2340-1120	2220-1040
KH6, Oahu, Hawaii	0400-1500	0400-1320	0500-1200
KL7, Anchorage	0230-1530	0200-1320	0200-1200
KZ5, Canal Zone	0100-1140	0000-1140	2320-1100
OX, Thule, Greenland	0140-1330	0000-1220	2200-1100
PY, Belem, Brazil	0100-0930	2340-0900	2200-0920
ZK1, Cook Islands	0530-1440	0500-1300	0800-1200

table 3. Predicted 15-meter openings.

to from	San Francisco	Dallas	New York
UAØ, Sakhalin Island	2040-0400	2120-0300	2200-0010
JT1, Ulan Bator	2300-0340	0000-0220	None
UA, Astrokhan	None	1440-1600	1220-1640
UA, Leningrad	None	1400-1600	1220-1740
DL, Berlin	1600-1800	1330-1900	1140-1910
VU2, New Delhi	0130-0320	1430	1240-1440
VS1, Singapore	2240-0440	2320-0300	1420-2000
	1730-1900	1620-1930	
DU, Manila	2200-0440	2200-0300	2240-2320
KG6, Guam	2000-0500	2010-0310	2020-0120
VK6, Perth	2100-0530	2130-0350	2130-0140
		1530-1830	
VK3, Melbourne	1940-0550	1930-0420	1940-0220
			1410-1540
VR2, Fiji Islands	1720-0550	1700-0420	1630-0220
		1500-1520	1410-1520
FO8, Tahiti	1600-0630	1530-0500	1530-0220
CE2, Valparaiso	1400-0340	1240-0440	1130-0120
PY1, Rio De Janeiro	1340-0300	1200-0200	1050-0040
ZD8, Ascension Island	1330-0220	1140-0100	1020-2340
CR6, Angola	1400-0200	1140-0040	1010-2300
ZS1, Capetown	1400-2200	1140-2220	1010-2320
5R8, Tananarive, Malagasy	1600-0240	1210-0020	1020-2300

table 4. Predicted 20-meter openings.

to from	San Francisco	Dallas	New York
UAØ, Sakhalin Island	1930-0540	2000-0500	2020-0300
JT1, Ulan Bator	2140-0510	2240-0440	2320-0300
			1200-1240
UA, Astrokhan	1530-1700	1240-1730	1100-1800
UA, Leningrad	1500-1900	1240-2030	1100-2010
DL, Berlin	1430-2030	1220-1230	1030-2100
VU2, New Delhi	0130-0320	1430	1240-1440
	1540	1330-1600	1120-1620
VS1, Singapore	2150-0630	1500-0440	1200-0220
	1630-2030		
DU, Manila	2200-0440	2200-0300	2240-2320
	1630-1830	1500-1800	1330-1800
KG6, Guam	1900-0650	1910-0520	1940-0350
	1610-1620	1450-1610	1400-1620
VK6, Perth	1600-2140*	1440-2000*	1400-0400
	0200-0730	0200-0610	
VK3, Melbourne	1530-2000*	1400-1900*	1250-1830*
	0120-0900	0000-1050	0030-0430
VR2, Fiji Islands	1530-0900	1400-1050	1250-1840*
			2200-0430
FO8, Tahiti	1500-1140	1340-1100	1240-0440
			0700-0900
CE2, Valparaiso	1300-1030	1200-0900	1020-0800
PY1, Rio De Janeiro	1240-0600	1050-0830	0940-0720
ZD8, Ascension Island	1240-0430	1030-0800	0920-0200
			0440-0620
CR6, Angola	1300-0430	1030-0600	0920-0130
			0300-0530
ZS1, Capetown	1830-0430•	1030-1240*	0920-1140*
		1740-0230	1620-0130
		0520-0800	0300-0600
5R8, Tananarive, Malagasy	1420-0420	1400-0500•	0920-0120
			0300-0400

* Between this and next time period, path may be open for 10-dB additional circuit gain or via long path.

• With 10-dB additional circuit gain, San Francisco-Capetown path open as early as 1240 GMT; Dallas-Tananarive open as early as 1110 GMT.

percent of the days of the month) from mid-latitudes (38° N.) to the south of east and west, with more frequent openings to latitudes below 33° N. Transequatorial forward scatter during evening hours may furnish your best chance to work 50 MHz DX. The TE season should open on February 10 at 8 PM local time.

Late word on 50 MHz beacon frequencies and liaison schedules may be obtained by sending me a self-addressed stamped postcard.

ham radio

tabla 5. Predicted 40-meter path openings during the ARRL DX contest.

to from	San Francisco	Dallas	New York
UAØ, Sakhalin Island	0500-1740	0500-1520	0500-1420
JT1, Ulan Bator	0720-1740	0800-1500	0800-1330
			2020-0100
UA, Leningrad	2300-1530	2200-1020	1900-0920
DL, Berlin	2240-0900	2140-0900	1900-0900
VU2, New Delhi	1120-1700	1200-1400	1920-0240
		2300-0220	
VS1, Singapore	1000-1740	1000-1540	1930-2330
DU, Manila	0820-1740	0820-1540	0900-1500
KG6, Guam	0640-1740	0700-1540	0700-1430
VK6, Perth	0930-1620	0930-1500	1000-1340
VK3, Melbourne	0800-1600	0800-1420	0800-1320
VR2, Fiji Islands	0500-1600	0500-1440	0520-1330
FØ8, Tahiti	0320-1520	0300-1420	0300-1320
CE2, Valparaiso	0040-1120	2340-1100	2240-1030
PY1, Rio De Janeiro	2340-0940	2300-1000	2130-0930
ZD8, Ascension Island	2330-0800	2200-0800	2040-0800
CR6, Angola	2300-0600	2200-0600	2000-0620
ZS1, Capetown	0000-0520	2230-0520	2030-0520
5R8, Tananarive	2300-0320	2200-0340	1940-0340

table 6. Predicted 10-meter openings.

to from	San Francisco	Dallas	New York
UAØ, Sakhalin Island	2200-0300	2250-0100	None
JT1, Uen Bator	0030-0215	0030	None
UA, Astrokhan	None	1430	1350-1440
UA, Leningrad	None	1430	1350-1440
DL, Berlin	None	1430-1630	1320-1700
VU2, New Delhi	0200	None	1340
VS1, Singapore	2320-0300	0000-0040	None
DU, Manila	2230-0300	2250-0110	None
KG6, Guam	2040-0320	2050-0200	2100-2300
VK6, Perth	2240-0400	2250-0230	2250-2330
VK3, Melbourne	2200-0420	2150-0250	2310-0100
VR2, Fiji Islands	1840-0420	1810-0250	1820-0100
FO8, Tahiti	1710-0430	1630-0300	1620-0100
CE2, Valparaiso	1450-0200	1320-0140	1210-2340
PY1, Rio De Janeiro	1430-0100	1240-0030	1140-2300
ZD8, Ascension Island	1440-0050	1220-2330	1110-2200
CR6, Angola	1510-0020	1230-2300	1120-2120
ZS1, Capetown	1430-2100	1220-2100	1110-2130
5R8, Tananarive, Malagasy	1600-2300	1310-2230	1130-2100

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low-noise, high-gain transistor preamplifier

An accessory for your receiver to improve s/n ratio from 80 through 6 meters . . . featuring the 2N2360 transistor

John J. Schultz, W2EEY, 40 Rossie Street, Mystic, Connecticut 06355

"Easy to construct" and "low cost" could very well be added to the title of this article. The preamplifier described is a solid-state device that will give excellent performance on any single band from 80 through 6 meters. Its size, construction, and cost allow it to be incorporated easily into any fixed station, mobile or portable receiver or transceiver, or to be used as a mast-mounted antenna preamplifier.

A good preamplifier can perform wonders in improving reception and reducing fading on bands where a receiver lacks gain, sensitivity, or has an image problem. This is true even with expensive receivers in single bands, because compromises must be made with receiver r-f stage design due to band-switching requirements and other constraints.

A good preamplifier should have several important characteristics, the main ones probably being a low noise figure and reasonable gain. Unless the preamplifier noise figure is less than that of the receiver's r-f stage, the preamplifier self-generated noise will mask signals that the receiver itself is capable of detecting. Even in the case where the preamplifier noise figure is not less than that of the receiver, the gain provided by the amplifier will still reduce the effects of fading and raise the level of all signals.

For best weak-signal reception, low noise

figure in a preamplifier is more important than gain, and most preamplifier designs sacrifice some gain to achieve low-noise performance. Reasonable selectivity to reject out-of-band signals and good overload response are secondary characteristics. If the receiver front end has good selectivity and overload response, then it's not necessary to incorporate these features in the preamplifier, thus keeping it simple.

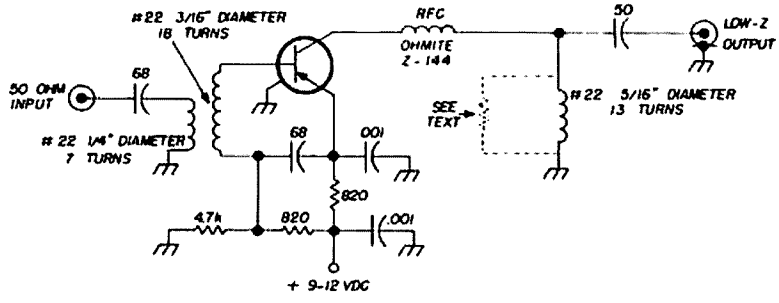
The preamplifier can be easily wired to be switched out of the antenna circuit if the preamplifier becomes saturated by strong signals. Achieving good overload performance using a simple circuit and a transistor is particularly difficult, although rarely does any problem result with the signal levels encountered in most amateur bands. Trying to use a sensitive preamplifier with a kilowatt transmitter operating a few block away, however, is quite another problem.

the 2N2360 transistor

The heart of the preamplifier is the Sprague 2N2360. This transistor has been proved in vhf-TV circuits, and although it may not seem as striking as an FET, it performs very well, allows very simple circuitry, and has recently become reasonably priced. (It formerly cost between $1.40 and $2.40 from most Sprague distributors.) If the 2N2360 is not available locally, a "substitute" type transistor should **not** be used, because its high-frequency performance will rarely match that of the 2N2360 as far as gain and noise figure are connected.

The 2N2360 will provide a 4- to 7-dB noise figure on any band from 80 through 6 meters. The gain will vary from about 15 to 25 dB or higher if tuned instead of broadband input and output circuits are used. The circuit used here requires a minimum of parts, and the entire preamplifier can be constructed for the cost of the 2N2360 alone if a "junk-box" selection is available.

fig. 1. Broadband preamplifier. Although coil values shown here are for 10 meters, basic circuit may be used on any band from 80 to 6 meters.

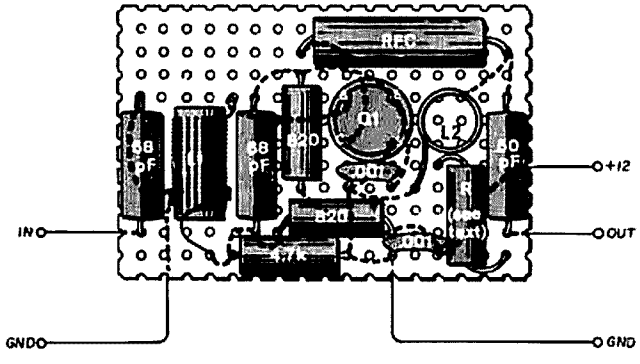


circuit

The complete preamplifier circuit is shown in fig. **1.** The basic circuit uses a grounded-emitter configuration. The input is double series tuned. A series capacitor matches the low impedance of a coaxial line. The secondary side is similarly tuned to match the 2N2360 input. As shown in **fig. 1,** the input circuit tunes to the high-frequency end of the band, and the secondary circuit tunes to the low-frequency end (the opposite condition could just as well be used).

Circuit resonance is easily checked with a grid-dip meter or a signal generator and receiver. Variable capacitors may be used initially to determine the necessary values and later replaced by fixed capacitors. These capacitors should have a good Q. Dipped silver micas (CDE type CD6 or similar) are recom-

fig. 2. Layout of the broadband preamp.



mended. Disc ceramics designed for bypassing should not be used, because their low Q will result in a loss of gain and selectivity.

The bias conditions, determined by the two 820-ohm and 4.7k-ohm resistors, were chosen as the best compromise between gain and noise figure. The 4.7k-ohm resistor can be varied if you want to experiment with the circuit or provide a manual gain control.

The parasitic choke in the 2N2360 collector lead prevents amplifier instabilty. The output circuit impedance is about 200 ohms and works quite well with simple capacitive coupling to the output coaxial line. If a more exact match is desired, a link of 2 or 3 turns can be used over the output coil. The output coil is resonated in the center of a band by using a grid-dip meter and adjusting the coil dimensions.

No loading resistor is used for operation on 80 through 15 meters and the phone portion of 10 meters. A loading resistor of about 1000 ohms is needed for complete 10- or 6-meter coverage. The only power required is from 9 to 12 volts. A battery, a simple rectifier circuit off the heater line in a tube-type receiver or any other well-filtered source may be used. Required current is less than 10 mA.

construction

A small Vector board is the handiest base for construction. With care, the preamplifier can be built to occupy about 1 inch by 1½ inch, which is small enough to fit into any receiver or transceiver. A suggested component arrangement is shown in **fig. 2.** This layout should be followed fairly closely, with particular care given to the separation and orientation of the coils to prevent intercoupling and oscillation. Normally, no instability or oscillation should exist. If this does occur, however, a shield across the middle of the 2N2360, connected to the case lead of the transistor, will solve the problem.

No special precautions are necessary if the preamplifier is installed inside an equipment enclosure, except that the ground connection should be as short and as direct as possible. If the preamplifier is used in a separate box (such as a minibox) as a mast-mounted unit, a 1-mH rf choke should be installed from the input coaxial line center conductor to ground. This will prevent any high static voltage on an antenna from damaging the 2N2360.

The component layout of the completed universal high-gain transistor preamplifier is shown in **fig. 2.** (Note particularly that the output coil is mounted **vertically** to prevent coupling to the other inductors). The construction shown may not deserve any kudos for neatness, but it does illustrate the ease with which the preamplifier components can be assembled.

in summary

Considering ease of construction and cost versus performance, the 2N2360 preamplifier circuit is certainly one of the best. It should be possible for the average amateur to build this preamplifier in one evening for any selected band with the aid of only a grid-dip oscillator. The improvement in performance that this preamplifier provides when used with most receivers and transceivers, particularly on 15 through 6 meters, is truly amazing.

I have presented only the simplest constructional use of the 2N2360 transistor preamplifier. Many variations are possible for those who would like to construct a more elaborate unit. Discrete tuning of the amplifier instead of the broadband operation, for instance, will increase receiver gain somewhat and improve image rejection. When used as a tuned amplifier, the input circuit series capacitor is omitted, and the series capacitor in the input circuit secondary is variable. The primary circuit serves, then, as a simple coupling link. The output coil can be either broadbanded or tuned by a variable capacitor.

If you want to construct a multiband preamplifier, there is no reason why the input and output circuits cannot be bandswitched. Provision should be made for disabling the preamplifier, if possible. Although it would take an unusually strong signal to overload the preamplifier, it is worthwhile to be able to determine positively, when signal distortion does occur, whether the preamplifier or another part of the receiving system is at fault.

ham radio

new products

fet principles, experiments and projects

Ed Noll, W3FQJ, noted for his previous works which include a three volume **Modern Communications Course** and handbooks for radiotelephone license exams, has done it again with this excellent and complete reference book on FET's.

This book covers the principles of operation of the field-effect transistor from the very beginning. While it includes chapters on theory and math for the engineer, it also explains how FET's work in everyday language for the experimenter and ham. It also shows applications as ac amplifiers, audio amplifiers and oscillators, and high-frequency amplifiers, oscillators and communications circuits. The projects are presented in an unusual way: they start with an audio amplifier, continue into detector and tuner projects and then combine them into one modern FET receiver.

Other projects include QRP transmitters, absorption wavemeters and field-strength indicators. All projects include complete parts lists and use parts available from almost any distributor; the experimenter won't be stymied by non-availability of needed parts.

This complete reference publication on field-effect transistors is a valuable addition to the library. $4.95 from your local distributor, or write to Howard W. Sams and Company, Inc., 4300 West 62nd Street, Indianapolis, Indiana 46268.

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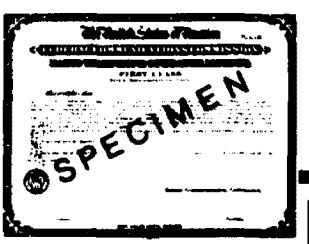
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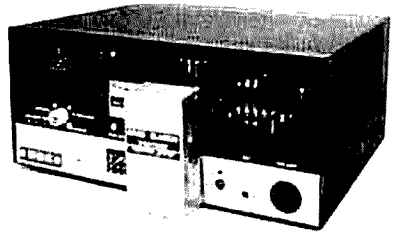
NOISE BLANKER

FOR THE SWAN 250

Westcom Engineering is now offering the TNB Noise Blanker in a version specifically designed for use with the Swan 250 transceiver.

The TNB-250 Noise Blanker effectively suppresses noise generated by auto ignitions, appliances, power lines, etc., permitting the recovery of weak DX and scatter signals normally lost in noise.

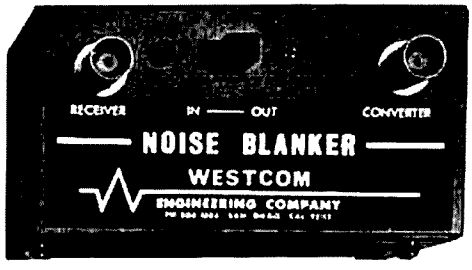
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TNB-250 shown installed on Swan 250 by means of the pre-punched accessory holes.

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Refer to the New Products column of the August '68 issue of Ham Radio Magazine for additional information on the TNB Noise Blanker or write for technical brochure.

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motorola power transistors

Motorola has just introduced the first six members of a new line of intermediate-power transistors. These new devices, in plastic packages, will dissipate up to 8 watts with suitable heat sinking. Although the package is only about the size of a TO-5 transistor, it can dissipate approximately 1 watt at an ambient temperature of 25° C.

The npn MPSUO1 and pnp MPSU51 are complementary transistors designed for use in audio amplifiers delivering up to 5 watts output. They have high gain and excellent current-gain linearity from 1 to 1000 mA. The maximum V_{CEO} rating for these devices is 30 volts, maximum continuous collector current is 1.5 amperes, and device dissipation is 8 watts.

The npn MPSU02 and pnp MPSU52 are intended for general purpose amplifier and driver applications. These silicon transistors have excellent gain linearity, low saturation voltage, minimum f_t of 150 MHz, maximum V_{CEO} rating of 40 volts and 6 watts dissipation. The npn MPSU03 and MPSU04 are 1-ampere devices for video output circuits and horizontal driver applications in TV sets. The collector-to-emitter breakdown voltage rating is 120 V for the MPSU03 and 180V for the MPSU4; total dissipation at 25° C case temperature is 8 W.

For more information on these new transistors, write to the Technical Information Center, Motorola Semiconductor Products, Inc., Box 20924, Phoenix, Arizona 85036.

in-circuit transistor tester

Amphenol has recently introduced a new compact transistor analyzer for checking transistors—in or out of a circuit—that also doubles as a very sensitive dc voltmeter. In addition, the unit functions as an accurate diode analyzer capable of measuring both forward and reverse current. This instrument was developed by Amphenol to minimize the time and effort required to pinpoint defective semiconductors. It can be used for troubleshooting all solid-state equipment, including receivers, hi-fi systems, TV sets, fm

tuners and tape recorders.

The new **Transistor Commander** can check high- and low-power npn and pnp power transistors for in-circuit dc-current-gain characteristics, check out-of-circuit devices for current gain, I_{CBO} and I_{CEO} leakage, check diodes and rectifiers for in-circuit opens and shorts and check out-of-circuit diodes for forward and reverse currents.

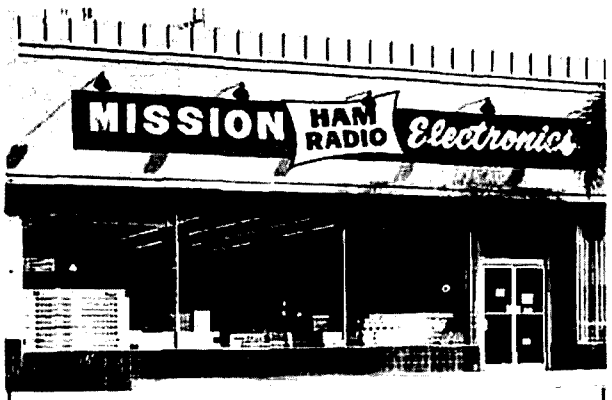
The new analyzer contains a current-limiting circuit for protection against accidental shorts; this prevents shorted transistors and diodes from damaging the instrument. In addition, simplified meter scales facilitate rapid testing, output test jacks are arranged so a quick-test socket can be easily adapted and the cabinet is constructed with a tilt leg for adjusting the meter to eye level.

The Transistor Commander provides a dc-current-gain range of 1 to 1000 (± 5%), adjustable collector current range of 0 to 10 mA and collector-base leakage and collector-emitter leakage measurements from 0 to 5 mA. In diode testing, a 100-mA limit circuit protects the diode from damage. The built-in 100-Vdc meter permits accurate measurements with the same probes that are used for the other checks.

For more information on the Model 830 Transistor Commander, write to the Amphenol Distributor Division, Amphenol Corporation, 2875 S. 25th Avenue, Broadview, Illinois 60153.

10 watts at 1000 mhz

Looking for high power solid state on the uhf bands? If so, here's a transistor that may interest you: the new S1050 from the Electronic Components Division of United Aircraft. The S1050 is a silicon npn epitaxial planar transistor designed specifically for high-power output in class-C uhf amplifier service. It uses an overlay emitter design and delivers 10 watts at 1000 MHz with 5-dB gain. The stud-mounted package is modified for optimum performance in high-frequency, high-power applications. Price, $200. For more informations, write to Electronic Components Division of United Aircraft, Trevose, Pennsylvania 19047.



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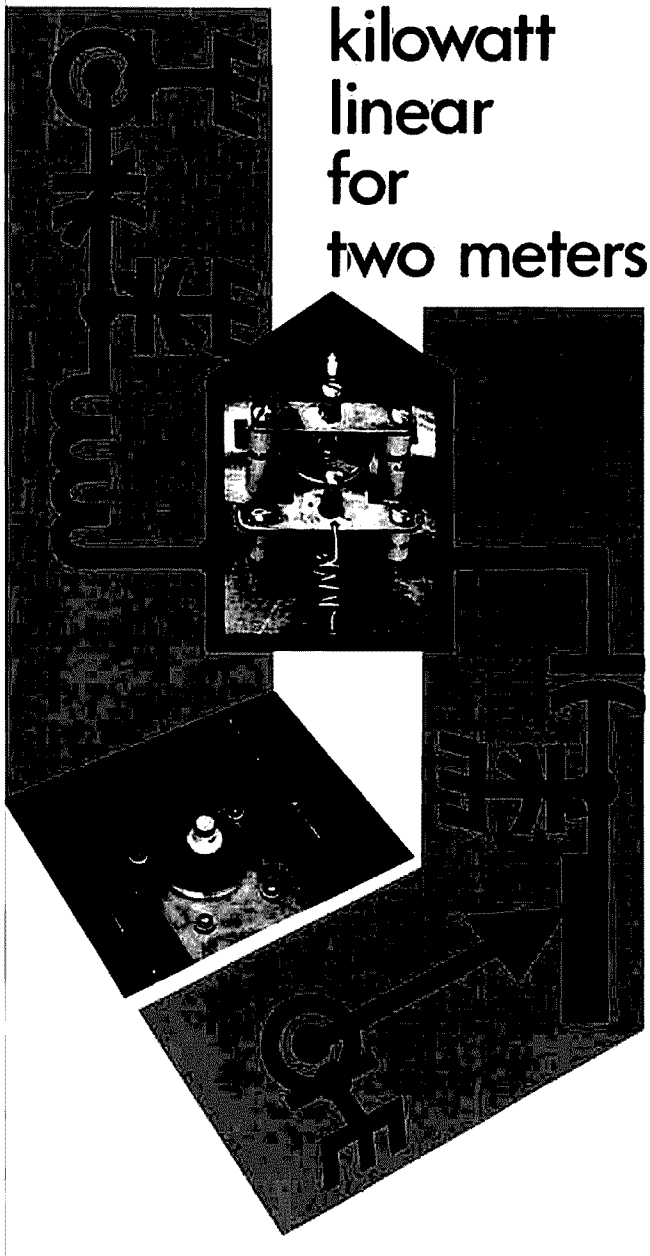
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MARCH 1969



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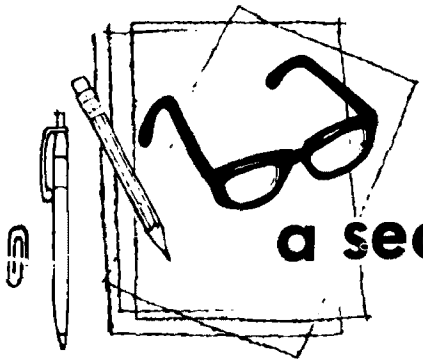
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a second look

by jim fisk

How many times in the last twelve months have you heard somebody comment to the effect that the days of creative basement and backyard engineering were over? With lasers, Gunn oscillators, fancy semiconductors and all the other esoteric paraphernalia that is announced every day by the scientific community, I would tend to agree. But occasionally there is a glimmer of hope—within the last two weeks I have heard of two cases where amateurs have made important contributions to the rest of the electronics world.

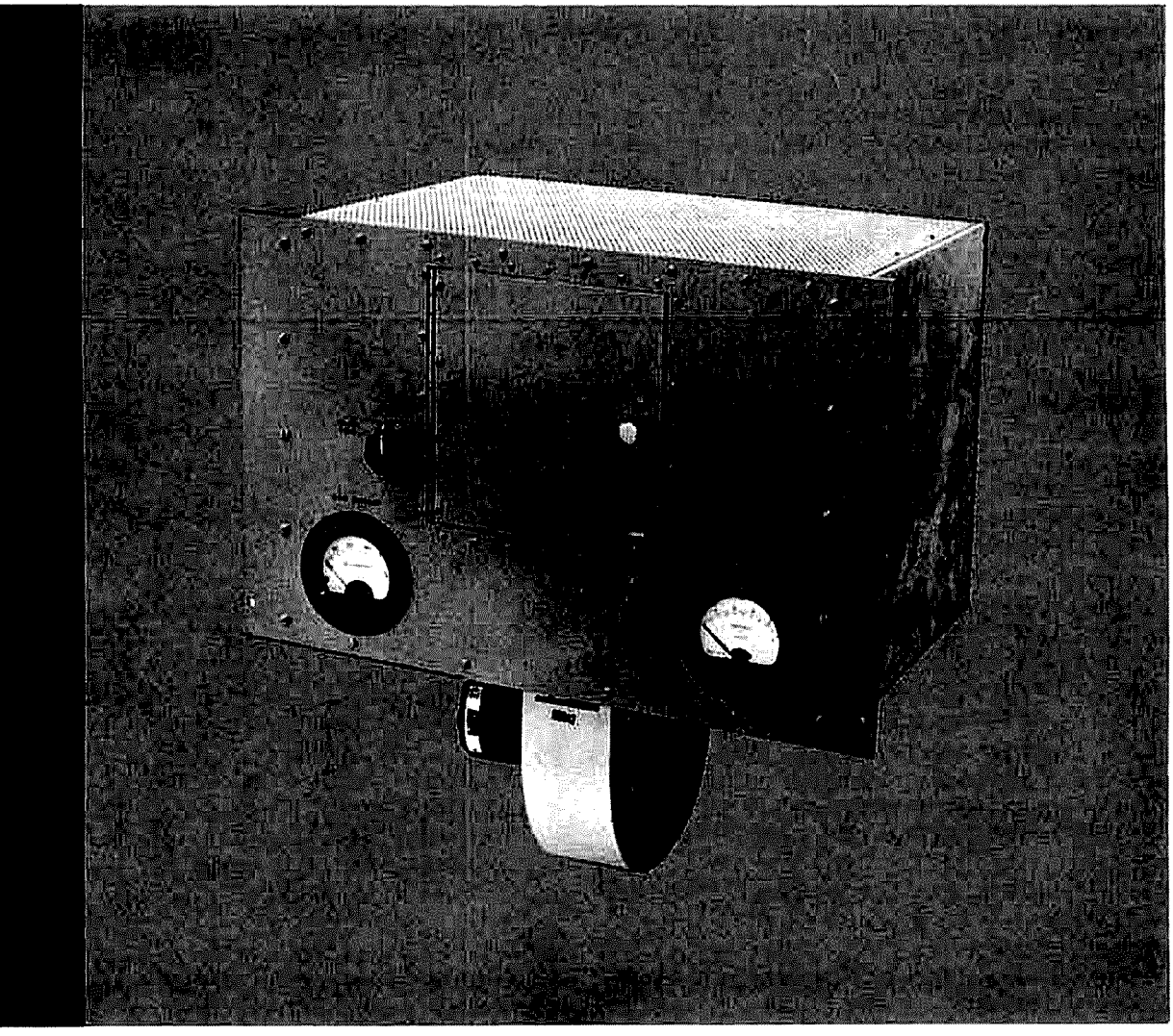
The first item concerns a Canadian, Syd Horne, VE3EGO, and an engineer friend, Doug Watson. Syd became interested in slow-scan television several years ago and was at the North American end of the first trans-Atlantic slow-scan television link. After experimenting with equipment originally designed by WAØNLQ, Syd recognized some of its limitations. Slow-scan requires eight seconds per frame, so any adjustments the operator makes in focus, brightness or contrast are difficult because it takes eight seconds to see the results; also, slow-scan vidicons are expensive. Doug Watson, after some prodding from Syd, came up with a solid-state camera and monitor which uses a fast-scan vidicon with sampling techniques to provide a slow-scan output. With this system, any adjustments are immediately visible and pictures are brighter and of higher quality.

The other development I heard of recently involves a new family of antennas that have performance characteristics that are pretty extraordinary—simple construction, switchable vertical or horizontal polarization, wideband and apparently, equal-to-rotary-beam performance. These antennas are the invention of George Bonadio, W2WLR, and are the result of backyard engineering work going back to 1953. At that time he was using a Windom antenna and wanted to broadband it. He took two Windoms, put them back to back, flared out the ends for broadbandedness, and found they presented a very broad reactance curve.

Experiments with different feeds, antenna tuners and element arrangements resulted in a family of four antennas with performance characteristics that apparently cannot be explained with the normal rules. A patent has been issued for the first antenna—the square-diagonal—and patents are pending on two and applied for on the last.

Some details of these exciting new antennas will be featured in next month's issue of **ham radio** along with the new sampling sstv camera and monitor from Canada. Although the slow-scan equipment was not completely amateur developed, since most of the design work was done in an industrial R & D lab, much of the ground work and certainly the concept stemmed from amateur slow-scan work. On the other hand, the family of Bonadio antennas was completely amateur conceived and amateur developed.

Jim Fisk, W1DTY
Editor



design data for a two-kilowatt vhf linear

The input matching network and plate circuit are unique in this strip-line amplifier—self neutralization is also featured

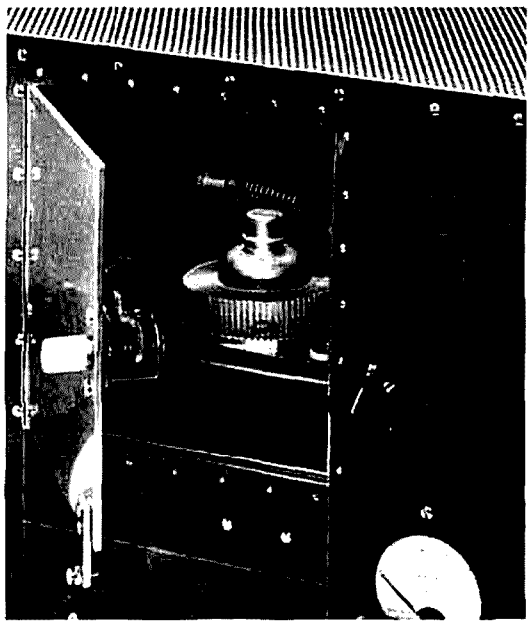
Robert I. Sutherland, W6UOV, Power Grid Development Lab, Eimac Division of Varian, San Carlos, California 94070

This article describes the design of a 150-MHz, 2000-watt PEP input, grounded-grid linear amplifier. I built this amplifier to conduct life tests and determine the performance characteristics of the new Eimac 3CX1000A7 high-mu ceramic triode at 150 MHz. Many of the design techniques and performance figures at 150 MHz will be of interest to the serious vhf and uhf experimenter. By changing the cathode matching network and the plate line length, this amplifier would be excellent for the 144-148 MHz band.

Several propagation modes, in which the maximum legal input power can be justified, are possible in the two-meter band. Meteor scatter, moonbounce, forward scatter and tropospheric bending are but a few.

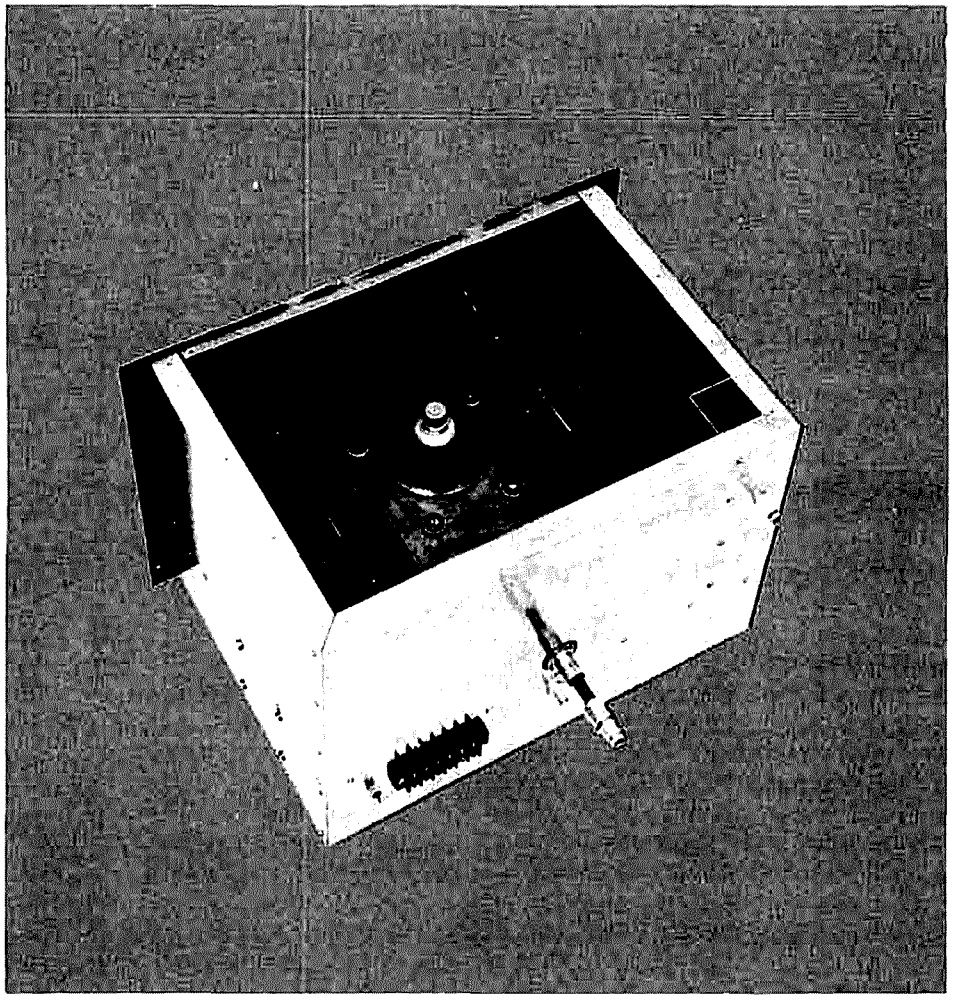
Running the full, legal one-kilowatt aver-

Amplifier with the access door open; the 3CX1000A7 can be seen mounted in the plate line.



age power input to the final amplifier in the cw and ssb mode has been fairly easy in the region between 2 and 30 MHz with the large selection of power tubes and circuits available. At these frequencies, many small tubes can be paralleled to develop

also use four such tubes to develop an average plate input power of 1 kW for ssb radiotelephone service, with reasonably low intermodulation distortion. A single 4CX1000K can be used to give 1 kW average input power, under voice conditions,



Top view of the amplifier with the shield removed. The variable tuning capacitor on the left was changed in the final model to a fixed capacitor. The output coupler coax may be seen protruding from the back panel.

power-handling capability. At 144 MHz, however, tubes that will handle 1-kW average power with good efficiency, stability and reliability are few, and large arrays of television sweep tubes are impractical.

choice of tubes

Many successful amateur designs have been built for 144-MHz, 1-kW continuous duty (single-tone) using the 4X150A, 4CX250A and 4CX300A tetrodes. Generally, two tubes are used in push-pull. You can

in a 144-MHz amplifier[1] with good efficency, stability and low drive.

Many popular high-frequency amplifiers, using zero-bias glass triodes such as the 3-400Z, 3-500Z, and 3-1000Z, are now being used in grounded grid. The 3-400Z and 3-1000Z have been used successfully in 150-MHz grounded-grid amplifiers, but neutralization is difficult.[2] The 3CX1000A7, a zero-bias, ceramic, high-mu triode is now available, which simplifies these problems. It is well-suited for vhf.

The 3CX1000A7 is electrically similar to the 3-1000Z. The amplifier described in the following paragraphs uses this metal-ceramic tube in a circuit for 150 MHz, and it has all the advantages of zero bias and grounded grid. Bias and screen-voltage supplies aren't required. Stage gain is high, and neutralization is simple and effective.

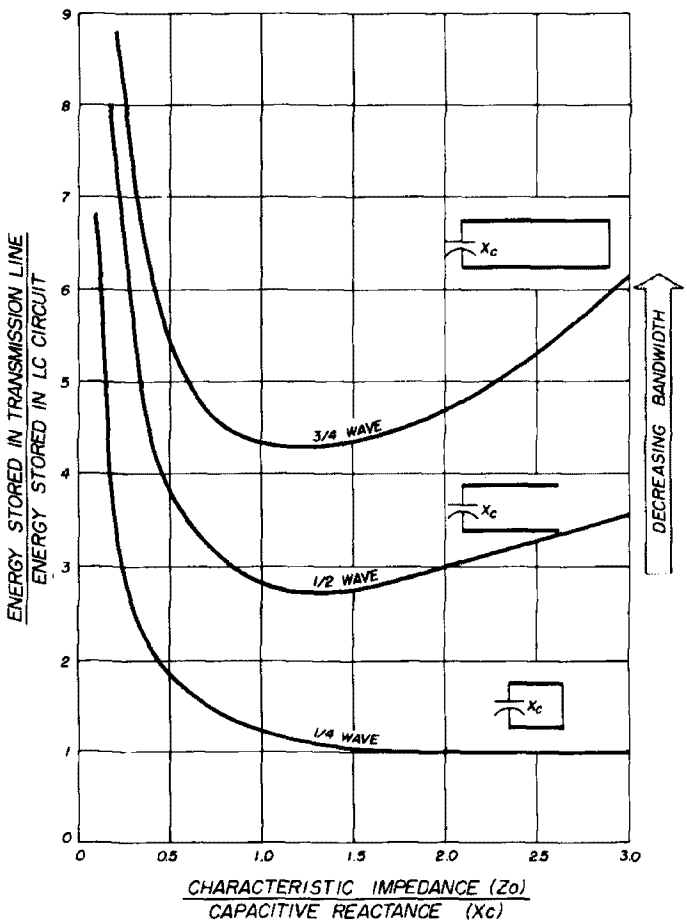


fig. 1. Frequency dependence of tuned transmission-line segments as compared to an LC circuit. Zo is the characteristic impedance of the line and Xc is the capacitive reactance acting on the sending end of the line.

the plate circuit

The choice of the plate circuit is important in the design of a vhf amplifier. I used a strip-line amplifier because it requires less machining.[3] A brake, shear and drill are all that are required. These tools are found in any well-equipped metal shop. You could do the job yourself with a hacksaw, file, drill and some perseverance.

The plate tank has a characteristic impedance, Z_o, of 60 ohms. The higher the impedance of this part of the circuit, the less will be the Q in the line; consequently, you will have a greater bandwidth. A wide bandwith isn't necessary to pass the modulation frequencies, but it is desirable if you want to eliminate a trimming device. A large amount of heat must be dissipated by the plate circuit, and if the circuit Q is low, the tank circuit will not detune as rapidly as it would with a high-Q circuit.

There's a practical limit as to how high you can make the plate strip-line impedance. Tube output capacity, plus strays, will limit the impedance and length of plate line:

$$X_c = Z_o \tan L$$

where:

X_c = tube output reactance.
Z_o = characteristic impedance of plate line.
L = line length in electrical degrees.

For a quarter-wave line, $Z_o/X_c = 2$ is adequate (**fig. 1**).

The plate strip-line is also useful for making a sandwich-type blocking capac-

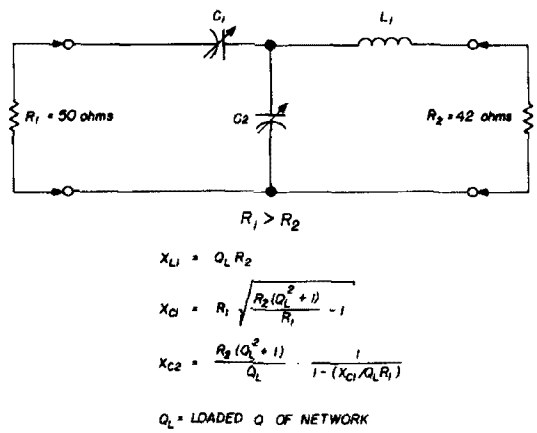


fig. 2. The schematic and equations used to calculate the input matching network. For this amplifier, R1 = 50 ohms, R2 = 42 ohms and assigned loaded Q is 2.

itor. In this strip-line assembly, a 10-mil-thick piece of **Isomica** was used for the dielectric. Mylar or teflon could be substituted with good results.

Output coupling is by a sliding contact on the inside surface of the plate line. As the sliding contact moves toward the anode, the area of the pickup loop increases, thereby increasing the coupling.

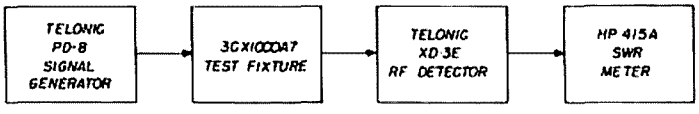
the input circuit

The input, or cathode tuned circuit, is a T network designed to match a 50-ohm coaxial line to the 42-ohm cathode driving impedance of the 3CX1000A7 (**fig. 2**).

The EIMAC SK-870 socket mounts directly on the chassis with the control grid contacts in contact with the chassis; that is, the control grid is at dc ground. The grid current is metered in the cathode return lead. The SK-870 socket is modified, as described later, to accomplish neutralization.

The driving impedance of the 3CX1000A7, in grounded-grid operation, is 42 ohms in parallel with about 20 pF of input capacitance. In order to simplify the design of the input circuit, it is assumed that the input to the 3CX1000A7 is resistive only. By making the matching network components variable, it's possible to resonate the input network, including the 20 pF of tube capacitance, while accomplishing the desired match.

fig. 3. Block diagram of the test set-up to determine the self-neutralizing frequency of the 3CX1000A7 triode and SK870 socket.



The input matching network in relation to the tube socket at the input coax line. The coil between the socket and the variable capacitor assembly is L1. The coil in the lower left-hand corner is the bifilar filament choke.

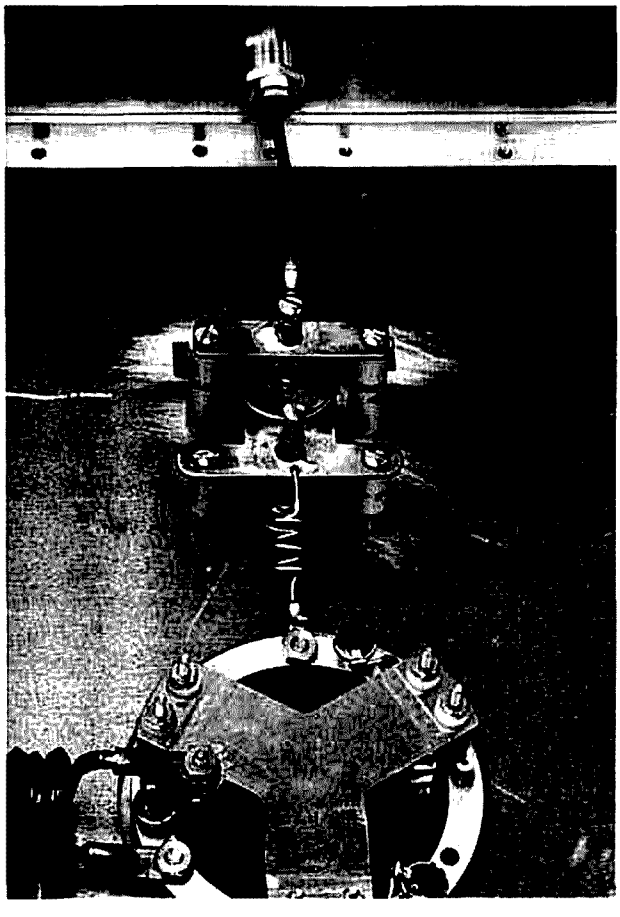


Fig. 2 shows the matching network and the design equations used. Under full plate-current load, the input match is adjusted to a 1:1 vswr, because only under full-load conditions will the tube input resistance be at the design value of 42 ohms. It's possible, however, to adjust the input network without the amplifier operating. Measure the resistance of a good-grade 42-ohm carbon resistor on an impedance bridge, then place the resistor between grid and cathode. The filament and all other voltages on the 3CX1000A7 must be off for this adjustment. A small amount of drive is applied through a standing-wave bridge. The input network is then adjusted for a 1:1 vswr at the operating frequency.

The "cold" adjustment won't be exactly correct, but the technique does allow the network to be tested and closely adjusted before application of power. The final adjustment is done with all voltages applied to the amplifier and tube with the 42-ohm carbon resistor removed.

self-neutralizing frequency

The 3CX1000A7 and SK-870 were tested in a special test fixture that treats the tube and socket as a four-terminal network. The anode was driven by a high power signal generator. The feedthrough signal was then detected on the cathode (see **fig. 3**). The control grid was grounded during this test. A curve was then plotted showing the relative magnitude of the feedthrough signal as a function of the signal-generator frequency.

The radio-frequency signal amplitude on the anode must be kept the same at all frequencies. A frequency will be found at which there is a minimum feedthrough signal; this point is called the **self-neutralizing frequency** of the tube and socket (**fig. 4**). The frequency thus obtained may be used as a guide in the design of a neutralizing circuit for the amplifier.

The graph of **fig. 4** indicates a self-neutralizing frequency of 105 MHz; this is lower than the operating frequency (150 MHz). The amplifier was unstable at 150 MHz before neutralization was applied. To raise the self-neutralizing frequency closer to 150 MHz, it was decided to lower the control-grid lead inductance. The SK-870 socket was modified by removing the six grounding spacers (see photo). These are metal spacers approximately 1/8-inch thick on each of the six SK-870 mounting screws. This modification allows the socket to be mounted 1/8-inch closer to the chassis.

The self-neutralization curve was again run with the same fixture and this modified

Tha Elmac SK870 socket. The six 1/8-inch matal spacers must be removed to neutralize the amplifier.

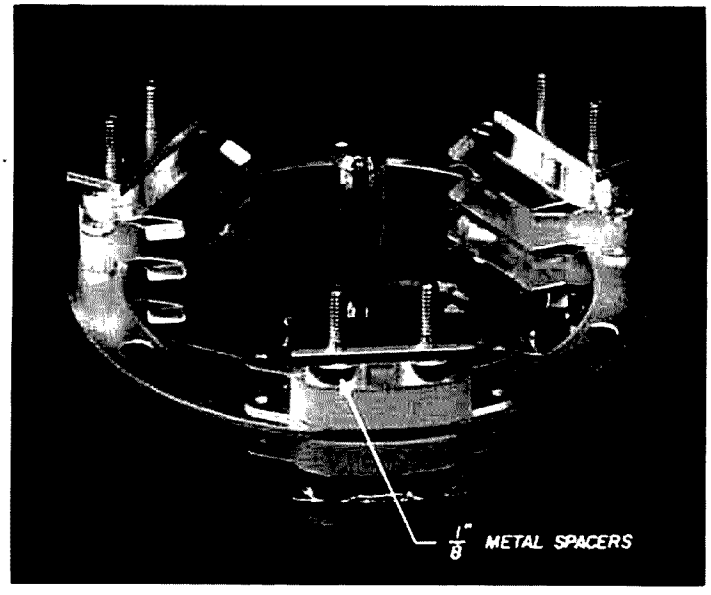


fig. 4. Relative input-to-output circuit isolation as a function of frequoncy before the socket modification. The self-neutralizing frequoncy is just below 105 MHz.

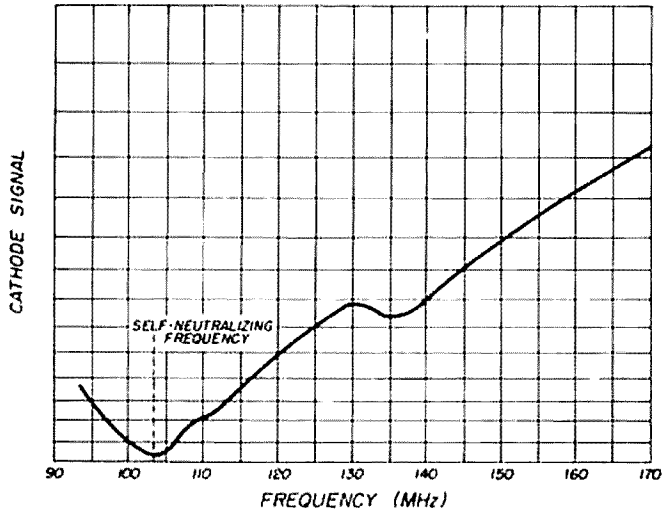


fig. 5. Relative input-to-output circuit isolation as a function of frequency after the sockot modification. The self-neutralization frequency has moved up to 136 MHz.

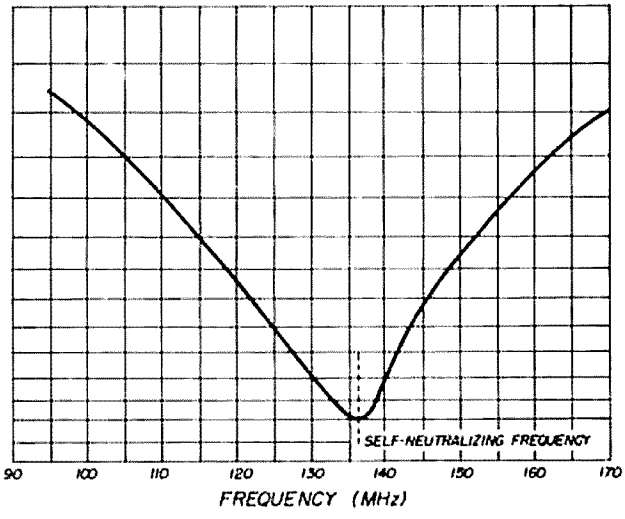
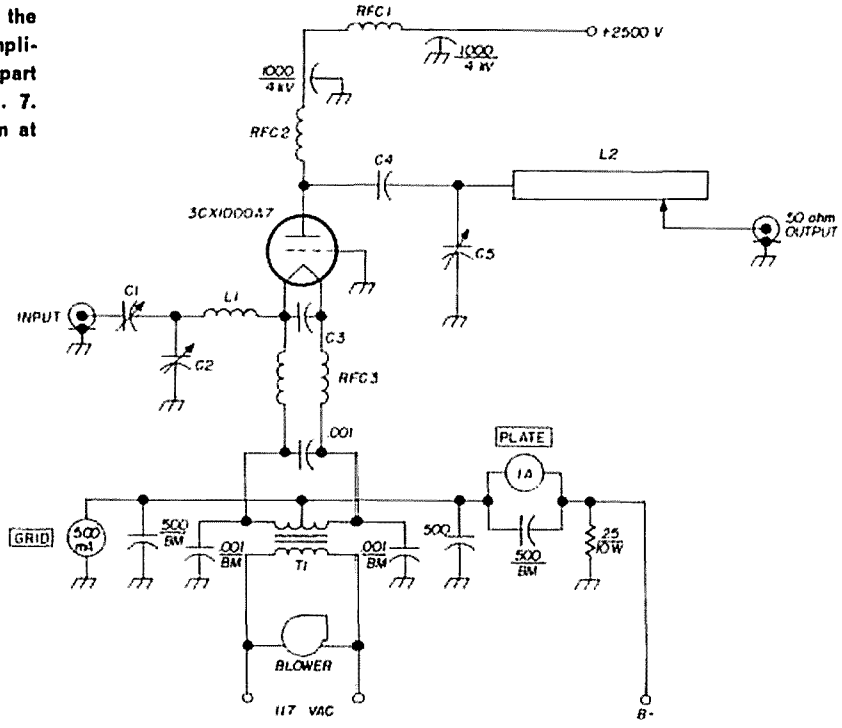


fig. 6. Schematic of the grounded-grid 150-MHz amplifier. C4, C5 and L2 are part of the plate line; see fig. 7. The blower delivers 37 cfm at 0.4 inch of water.



C1 11.9 pF (see matching network photo)

C2 1.06 pF (see matching network photo)

C3 1500 pF. Three 500-pF, 500-V stud-mounted button micas, spaced uniformly around the filament ring

L1 0.089 μH. 3 turns no. 14 wire, 1/16" diameter, 5/8" long

RFC1, 2 10 turns no. 16 tinned, 0.650" OD, 1-1/2" long

RFC3 bifilar coil no. 10 wire, each coil 5 turns, 3/4" ID, 1-5/8" long

T1 5 V, 33 ampere filament transformer

socket. Fig. 5 shows the results: the self-neutralization frequency has now been moved to 136 MHz. The socket was again mounted in the amplifier, which was then retested for stability. The amplifier was stable, and ihe power output peak coincided with the plate current dip. Either arrangement of the socket grid return leads raised the self-neutralization frequency sufficiently to achieve stability or the neutralization point is sufficiently broad to include 150 MHz. It wasn't necessary to improve ihe neutralization further. Undoubtedly this simple socket modification is necessary to achieve optimum stability in ihe two-meter band.

amplifier assembly

The schematic of the amplifier is shown in fig. 6. Note that the circuit is much the same as that of an hf amplifier. Only the

Closeup of the output load coupler. The teflon block, with the fingerstock mounted on top, forms an adjustable loop as it slides back and forth to adjust the loading.

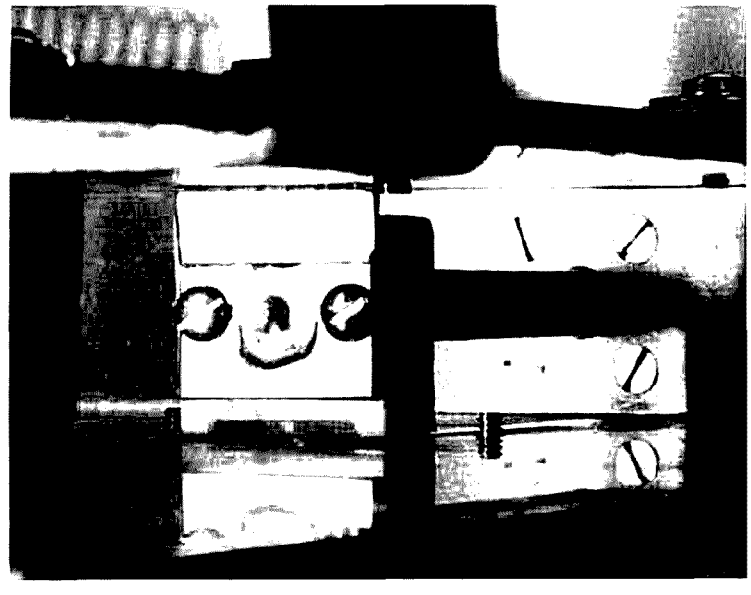


plate tank circuit and the input matching network can be considered unique. The metering of the grid and plate currents is in the negative return leads. The grid is metered in this manner to allow a very low impedance radio-frequency connection between the control grid and the chassis.

As noted previously, this grid circuit is part of the amplifier neutralization scheme, and the use of a grid by-passing arrangement could have complicated the amplifier neutralization. The plate current is metered in the negative return lead in the interest of safety. It is always dangerous to have a plate current meter mounted on the front panel if the meter is operating at a high potential with respect to ground.

A 25-ohm, 10-watt safety resistor must be connected between the negative terminal of the plate power supply and ground. If either grid or plate meter should become open circuited, and the plate side of the power supply shorts to ground, the negative side of the plate supply could assume a potential equal to the supply voltage. The 25-ohm resistor will prevent the negative terminal voltage from soaring and will load the power supply sufficiently to cause the fuses, or overloads, to function.

Fig. 7 is a pictorial of the plate line and the dimensions that were used in the 150-MHz amplifier.

operating results

The amplifier has been operating at 150-MHz many hours under the following conditions:

Plate voltage	2500 volts
Plate current	800 milliamperes
Plate input power	2000 watts
Grid voltage	0 volts
Grid current	240 milliamperes
Output power	1175 watts
Drive power	50 watts
Filament voltage	5 volts
Filament current	33 amperes

Note the output power and the required drive power; the power gain is 23.5 times or 13.7 dB.

No intermodulation distortion tests were made at 150 MHz although tests have been run at 2 MHz to determine the tube characteristics. Under the same operating conditions given for 150 MHz, the third-order intermodulation products were no less than 32 dB down, and the fifth-order products were better than 38 dB down from one tone of a two-equal-tone test signal.

references

1. Barber et al, "Modern Circuit Design for VHF Transmitters," CQ, November & December, 1965.
2. Orr and Sayer, "Semi- and Super-Cathode Driven Amplifiers, QST, June & July, 1967.
3. William I. Orr, W6SAI, and John T. Chambers, W6NLZ, "Strip-line Kilowatt for 432 MHz," *ham radio*, September, 1968.

ham radio

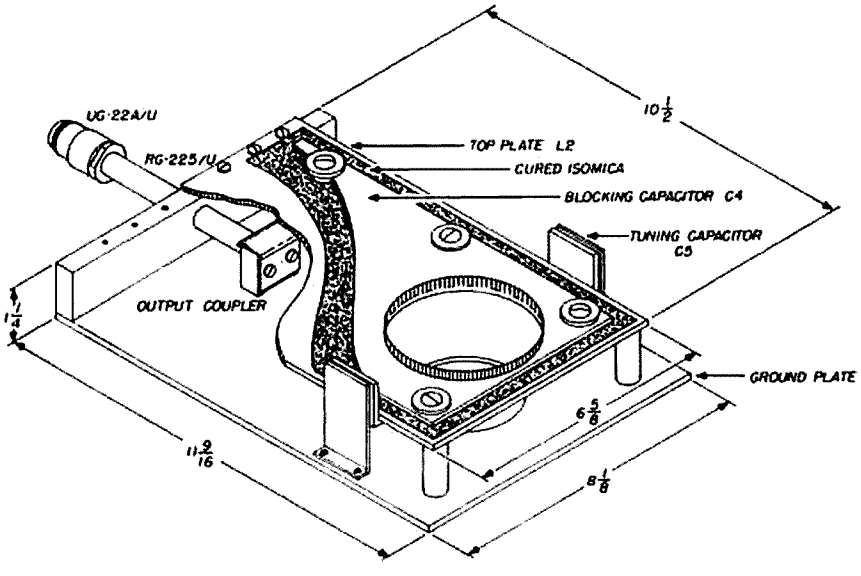
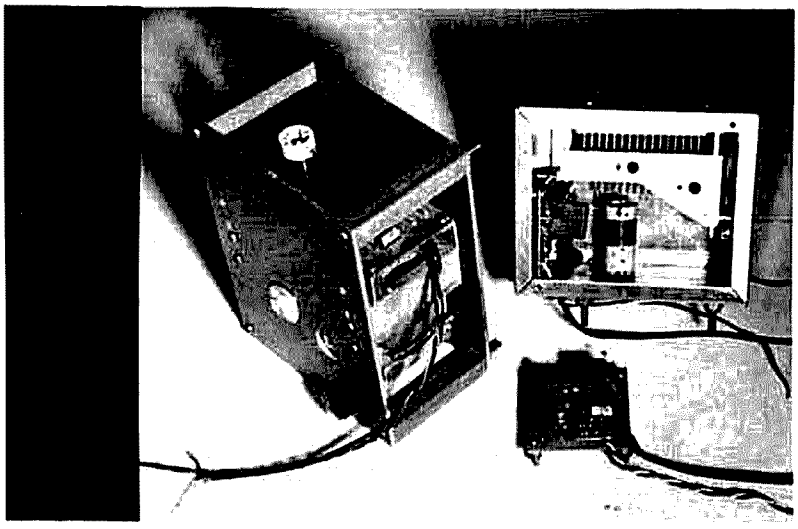


fig. 7. Cutaway view of the 3CX1000A7 plato line. The isomica is the dielectric for tthe plate-blocking capacitor C4. The two support pillars and six insulating shoulder washers are made of teflon as is the sliding output coupler block. The finger contact strip for the anode of the 3CX1000A7 is Instrument Specialties Company no. 97-133. A moro detailed drawing of the plate line is available from the author.



simple high-stability variable frequency oscillator

Some experimental and practical construction notes on a stable solid-state Seiler oscillator

Bill Wildenhein, W8YFB, RD 2, Blanche Avenue, Elyria, Ohio 44038

How would you homebrewers like a vfo with less drift than most factory-built ssb gear, very fast warmup, linear dial calibration, a capacity for using a wide variety of silicon or germanium transistors (pnp or npn) without any circuit changes, unusually flat output vs frequency, low cost, compactness and the capacity to be built with simple hand tools? So did I. After building and testing four different vfo circuits, I chose the circuit in **fig. 1** as the most reproducible vfo circuit with good operating characteristics.

circuit

The oscillator circuit I finally selected was W3JHR's (now W3JM) synthetic rock.[1] My first try with this circuit showed that Captain Lee's performance claims were very modest, and some of the circuit advantages weren't even mentioned. My next effort used the same circuit but was ruggedly built in a box of precision-machined hard aluminum with 1/8-inch end plates; this served as a "mechanical standard." This model (oscillator

number 1) used two RCA 2N247 germanium transistors.

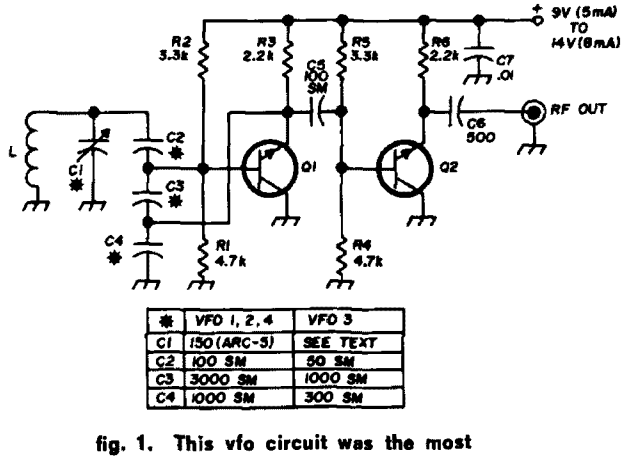
A second model (oscillator number 2) was built with plastic-encapsulated silicon transistors (dc beta about 50) in a 3x4x5-inch aluminum utility box; the layout and circuit board are shown in **figs. 2** and **3.** However, in my unit, the circuit board was not etched; component leads were used for all interconnections. When this much more simple construction was compared to the "mechanical standard," it was obvious that it was completely adequate for single-sideband. If it were included in a complete ssb package, of course, much of the total stability would depend upon the rigidity of its mounting on the main chassis.

The drift characteristics of oscillators number 1 and 2 are plotted in **fig. 4.** Note how closely the drift curves of the two units paralleled one another when the ambient temperature was deliberately forced up 10 degrees. This is a good indication of what will happen if you mount the completed vfo next to a well-stoked 6146! Temperature compensation is definitely in order for chassis position with wild temperature ranges.

The drift runs plotted in **fig. 5** were made with the oscillators covered by small cardboard boxes. Since the first hour of these runs was spent checking out the instrumentation, no warmup drift is shown. With such minimal protection against the temperature effects in my basement—during a particularly cold spell in January—I felt the performance was nothing short of remarkable.

The drift runs plotted in **fig. 6** were made with oscillator number 2 protected by a 1/2-inch thick styrofoam box. Warmup drift in curve A was quite reasonable, although a slight improvement was obtained (plotted in curve B) by replacing the metal tuning shaft with a bakelite shaft to minimize heat transfer into the styrofoam package.

The next vfo I built—oscillator number 3—was a try at further miniaturization; I also wanted to check the circuit at higher frequencies. The vfo was built into a small 4x4x2-inch aluminum utility box as shown in **fig. 7.** The feedback capacitors C3 and C4 used in oscillators number 1 and 2 at 3.5 MHz were much too large at 9 MHz; unless I used a transistor with extremely high gain, the circuit wouldn't oscillate. The values I finally used are shown in **fig. 1.**



*	VFO 1, 2, 4	VFO 3
C1	150 (ARC-5)	SEE TEXT
C2	100 SM	50 SM
C3	3000 SM	1000 SM
C4	1000 SM	300 SM

fig. 1. This vfo circuit was the most stable of several tried; it was originally developed by W3JHR.

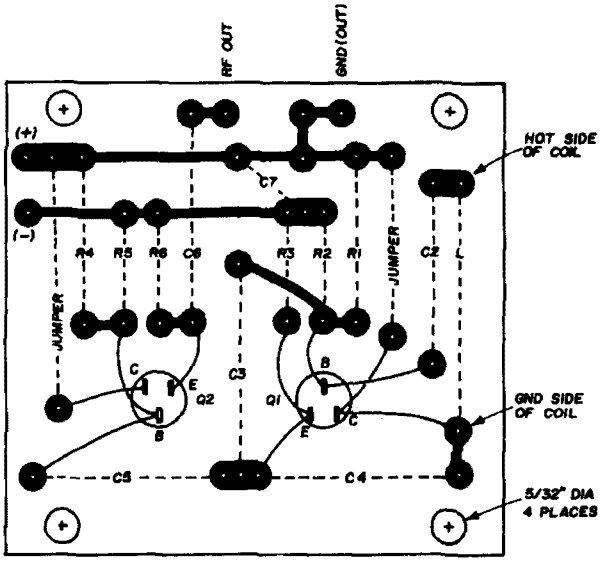


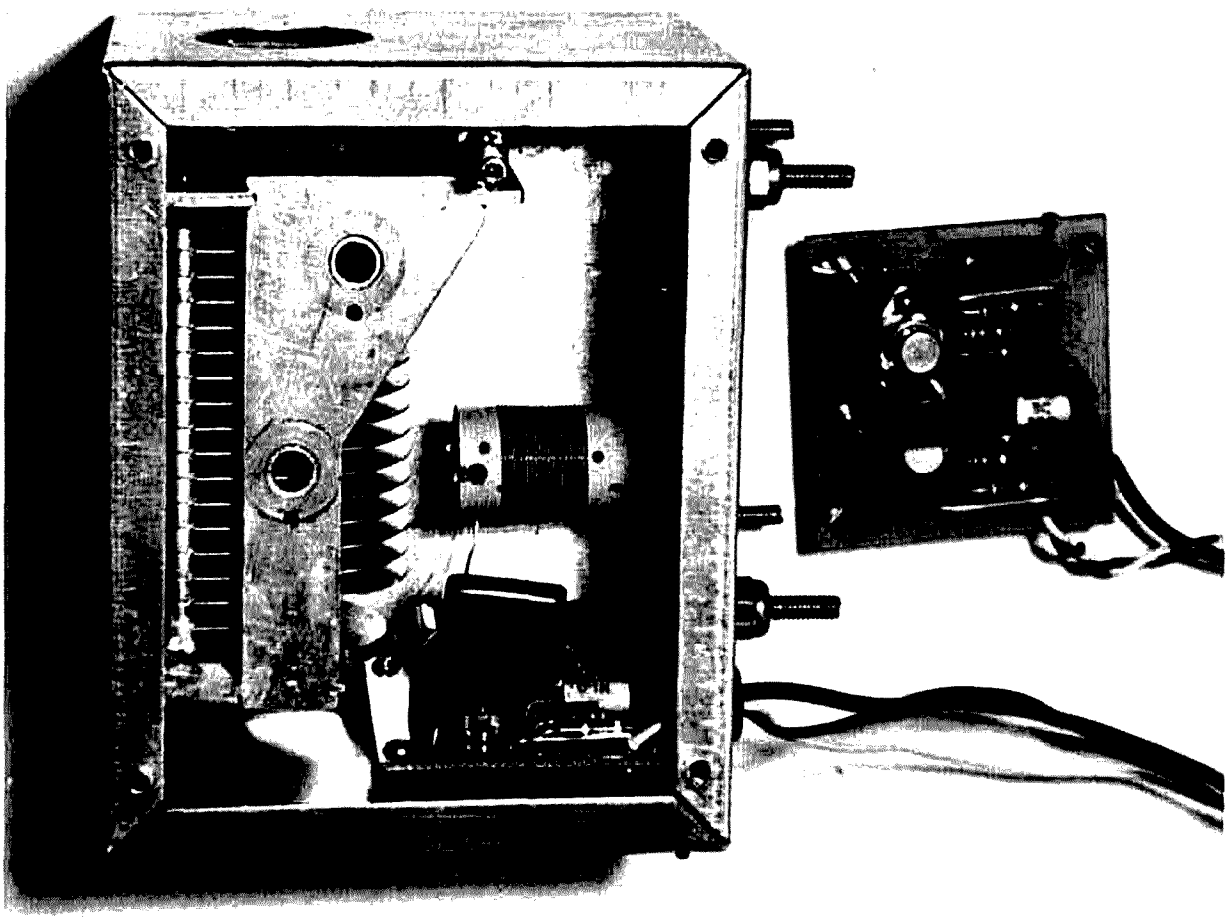
fig. 2. Full-size printed-circuit layout for the vfo shown in fig. 1.

The first variable capacitor I used in oscillator number 3 was a popular low-priced double-bearing 15-pF unit. It was a bit stiff to rotate, and vfo drift was terrible: jumps of 1 kHz in 5 minutes were common each time the dial was reset to a different frequency. This capacitor was replaced by a single-bearing 15-pF Millen 22915 variable. A single-

bearing variable is usually not mechanically stable where any vibration is present, but for bench tests it is adequate.

The new Millen capacitor operated very smoothly and required very little torque. The drift run plotted by curve A in **fig. 8** showed promise. The capacitor was then "exercised" by chucking the shaft in an electric drill and running it in a bit while feeding oil to the bearing. Several days later a couple of frequency runs showed that the rotation stic-

oscillator number 3 will make a useful vfo for vhf work. Another possibility for vhf work would be to put a larger capacitor across the feedback capacitor C4. If a 100-pF variable were placed in parallel with the 250-pF mica capacitor used for C4, the vfo will tune from approximately 8333 to 8500 kHz. This should make a very stable vfo for the low end of 6 meters. The change in feedback should not materially affect over-all stability, although the output voltage may vary somewhat from



tion had been largely eliminated: plot B in **fig. 8** is typical. Further improvements in mechanical stability could probably be obtained by replacing the utility box covers with covers made from 1/8-inch aluminum sheet.

If you have a very smooth capacitor that will fit into a small utility box, the layout of

one end of the band to the other.

Since oscillator number 2 seemed to represent an optimum type for general purpose construction, an identical model was built to check reproducibility. An ARC5 tuning capacitor was used but the oscillator was set up for oscillation at 9 MHz. The oscillator

transistor was a plastic encapsulated silicon device with dc beta between 400 and 500. As you can see from the drift characteristic plotted in curve B of **fig. 9,** it leaves something to be desired.

This same vfo was then shifted to 2 MHz. At 2 MHz oscillation was easily obtained with a pair of unmarked surplus silicon transistors. The frequency change was an experiment to see if such misfit semiconductors would have any significant effect on the drift characteristic. As you can see from curve A in **fig. 9,** the unit was quite stable.

The only adverse effect I could find with this circuit was the frequency shift per volt change in power supply voltage. With the misfit transistors, drift was 500 Hz per volt; all the other oscillators exhibited a shift of 100 to 125 Hz per volt. In one case, with oscillator number 1 at 2.5 MHz, the drift was 45 Hz per volt; at 4 MHz, this same oscillator showed 125 Hz per volt drift. This is not at

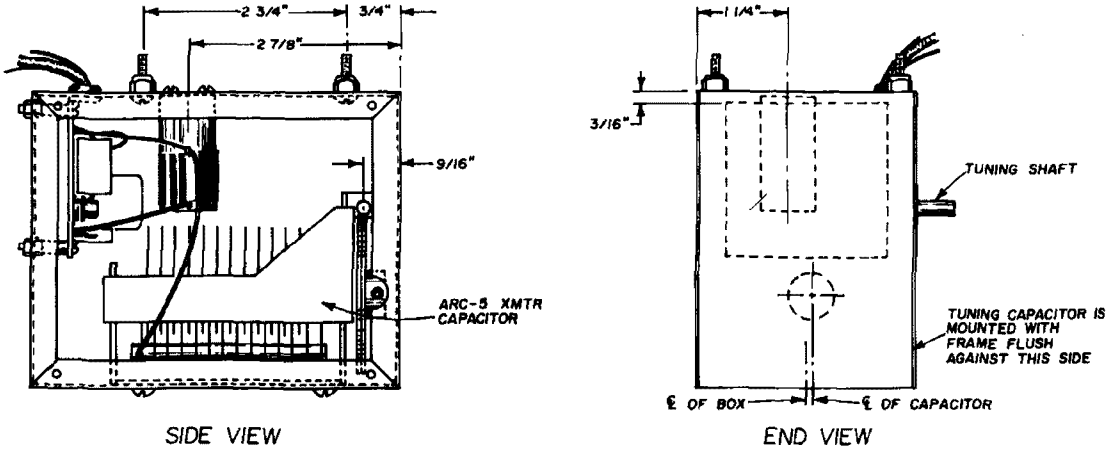


fig. 3. Mechanical construction of vfo number 2.

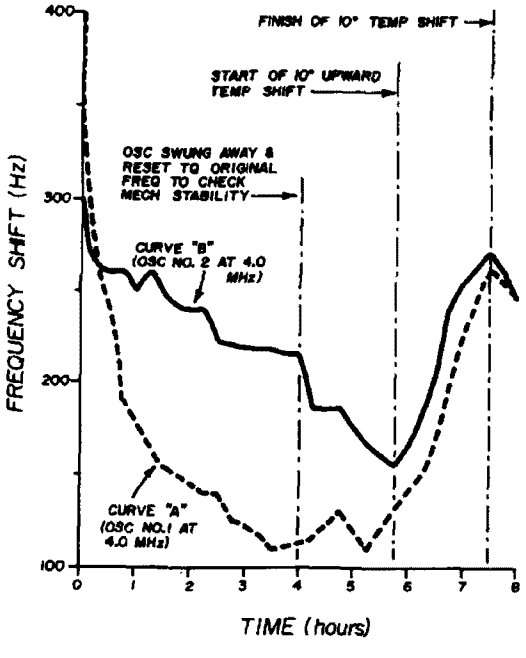


fig. 4. Drift characteristics of oscillators 1 and 2.

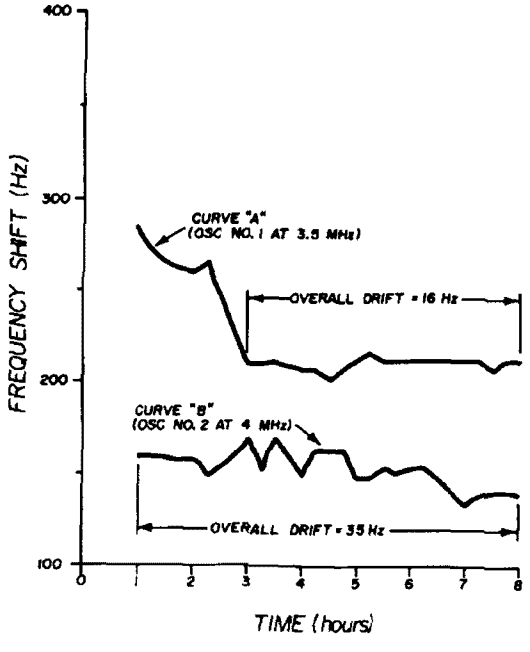


fig. 5. Drift characteristics with the vfo in a fairly constant temperature environment.

all severe and can be cured with a simple voltage regulation circuit.

If your application calls for a vfo output other than 3.5 or 9 MHz, it may be necessary to change the values of the coupling and feedback capacitors for optimum results. As a starter, the reactance of C2 should be about 600 ohms, C3 about 20 ohms and C4 about 60 ohms at the frequency of operation.

Since the output of the vfo itself is quite low—on the order of 0.5 to 1.5 volts peak to peak—it will be necessary to add a buffer amplifier for most applications. The output impedance of the oscillator stage is fairly high, so a high input impedance buffer stage using an fet or Darlington-connected transistor is required to minimize loading.

circuit advantages

Perhaps one of the most interesting advantages of this vfo circuit is the remarkably flat output voltage vs frequency characteristic.

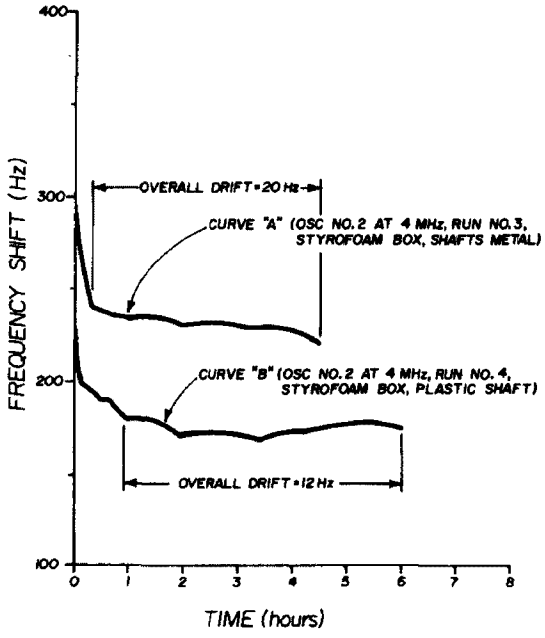


fig. 6. Drift characteristics when the vfo was protected with a ½-inch styrofoam box.

All of the oscillators were checked with a Tektronix scope, and output variations were 1 dB or less over all of the bands checked. This is obviously an advantage if the vfo is fed directly into a mixer stage. The mixer can be set up for optimum characteristics with assurance that they will not change with the vfo setting.

Another plus factor for this circuit is the clean waveform. Unless the feedback ratio—determined by C3 and C4—is too far off, or the reactance of the group of capacitors is too high (vfo on too low a frequency), waveform is excellent. This is an invaluable aid when designing a high performance mixer circuit.

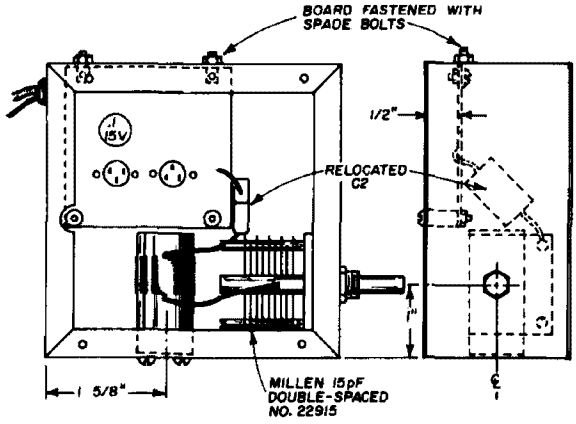


fig. 7. Mechanical construction of vfo number 3.

tank circuit components

I can't stress too highly that the real heart of these vfo's is the tuning capacitor and coil. It's probable that a good deal of the excessive drift on oscillator number 3 could be traced to one of these components. The variable capacitors used in oscillators 1 and 2 were originally installed in ARC5 transmitters. This capacitor is probably the best all-around choice for these vfo's; it is well built with high grade parts, and the gear drive ratio is adequate for the job. This is important because a drum dial can be attached to the end of the gear drive; this will save the cost of a separate dial and drive assembly. Another advantage of the ARC5 capacitor is dial calibration—it is essentially linear.

If you use a different type of capacitor, watch for erratic frequency jumps each time the capacitor is reset. To assess the merits of a particular capacitor in this respect, set the dial on a particular frequency from a clock-

wise direction. Make another run the next day, but approach the same frequency from the counter-clockwise direction. After about four frequency-setting runs you'll see the difference. This drift is mechanical and will usually be opposite to the direction you turned the capacitor.

The coil must be a rigid, quality unit. In one model I had good results with a high grade phenolic form, so this material isn't completely hopeless. I also tried commercial airwound coils; I found the best way of mounting these was to cement a block of polystyrene to one of the supporting ribs with Q-dope. Tests seem to indicate that if the leads from the coil to the circuit board are not preformed to fit with a minimum of stress, the built-in stress becomes a source of drift later on.

The oscillators used for the frequency drift runs presented here all used ceramic coil forms. The average homebrewer will have the best results with grooved forms if they're available. If they're not, try to get some phenolic tubing and have a machinist friend lightly groove several pieces at various pitches to give you some experimental forms.

In one vfo I built, I found a mysterious and sudden frequency shift. This was eventually traced to the manner in which the coil was terminated. The coil form had a hole near the end, and the coil wire was pulled through the hole several times to secure it. When the wire was pulled through the hole, the insulation was damaged and resulted in a partial tiny turn that occasionally shorted out. From there on, I used small bolts and nuts to form a stud to solder the wire to.

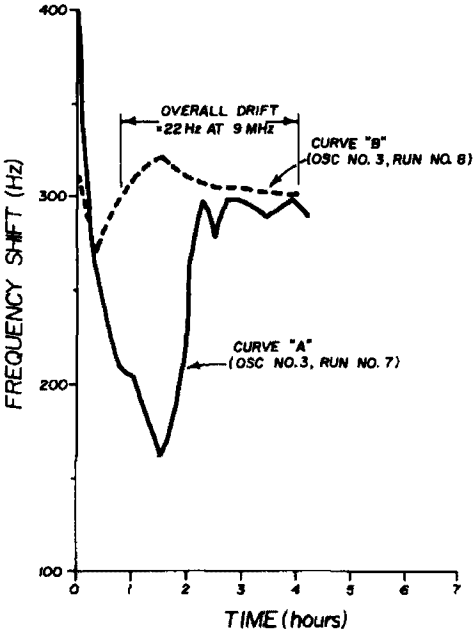


fig. 8. Drift characteristics of oscillator number 3.

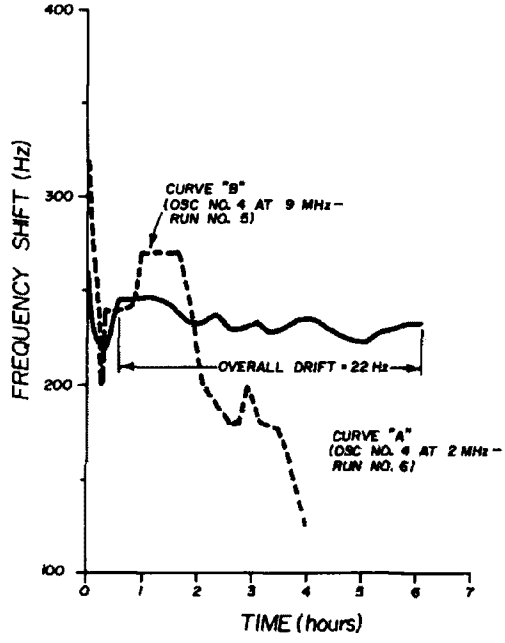


fig. 9. Drift characteristics of oscillator number 2 when it was shifted up to 9 MHz and down to 2 MHz.

Table 1 lists the approximate inductance required for any output frequency range be-

table 1. Tank circuit inductance for different frequency ranges.

vfo range (MHz)	inductance (μH)
1.7 - 2.15	34
2.0 - 2.5	25
2.5 - 3.1	16
3.3 - 4.2	9.5
4.0 - 5.0	6.3
5.0 - 6.2	4.0
6.0 - 7.5	2.8

tween 1.7 and 7.5 MHz with the ARC5 variable capacitor. No exact coil winding information is presented here because each individual builder will probably use a coil form available in his junk box. In the vfo's that I built, no effort was made to select form factor (coil length-to-diameter ratio), wire size or spacing for maximum Q. Several coils that I used departed a great deal from optimum, but results were uniformly good.

measurements

To make any kind of reasonable frequency measurements, you'll need a good crystal standard powered by a well regulated supply and capable of being zeroed in on WWV. A 100-kHz standard is generally poorer than a 1-MHz standard, but either will suffice if you use care. I use a BC-221 frequency meter with a regulated power supply in the "crystal only" mode. Owners of the LM version can shift the heterodyne oscillator out of the ballpark and depend on the crystal by itself. With a four-hour warmup, either model makes a good standard.

Many BC-221 owners are not aware that some models have a crystal frequency adjusting capacitor located underneath the nameplate on the front panel. When making a drift run, I secure the nameplate with masking tape to permit quick access for zeroing to WWV; it's best to replace the nameplate each time to minimize thermal shifts.

To measure vfo drift, the beat note between the 1-MHz standard and the vfo is compared to the audio frequency output of a Hewlett-Packard 200C audio generator. The HP-200C is very accurate and stable with a four-hour warmup period. Nevertheless, I compare the output to the 60-Hz line frequency for maximum accuracy. You can use any old oscilloscope for the Lissajous patterns.

A homebrew solid-state phase-shift audio generator could be built to cover 200 to 1000 Hz for this work. If calibrated at 60-Hz intervals, with the 30-Hz points estimated, it should be suitable for most purposes.

If you use a homebrew audio generator, it will be necessary to minimize error due to its drift. One way is to make all your frequency measurements at a multiple of the vfo frequency. For example, if a 3-MHz vfo and crystal standard are compared at 3 MHz and the vfo drifts 10 Hz, you must be able to rely on your audio interpolation oscillator to within 10 Hz. On the other hand, if you feed the output of the crystal standard and vfo into a receiver tuned to 30 MHz, you can measure the tenth harmonic.

Any drift will also be multiplied by ten, so a 10 Hz drift at 3 MHz becomes a 100 Hz drift at 30 MHz and is easily read on your audio oscillator comparison setup. Be careful: the crystal standard frequency is also multiplied! If it was set within 1 Hz of WWV at 5 MHz, 30 MHz is six times higher, so your crystal-standard error is 6 Hz.

Any oscillator will shift frequency with changes in supply voltage. If your vfo shifts 200 Hz with a 1 volt change in supply voltage, you must be able to measure voltage changes as small as 0.05 volt to be sure that frequency shift doesn't exceed 10 Hz.

One simple way of measuring small voltage changes is illustrated in **fig. 10.** In this circuit a zero-center microammeter compares the voltage difference between the power supply and a reference battery. As long as no difference in voltage exists between the power supply and the reference source, no current flows through the microammeter. With this setup, you can set the power supply voltage within about 0.01 volt.

Since the battery operates under no load, it is a stable reference if it's in good condition. The power supply must be pretty well regulated or you will spend all your time following its variations trying to keep the meter zeroed.

If you want to set your power supply to other voltages with the same precision, the reference must be changed. A string of fresh flashlight batteries connected in series provides a simple means of setting up a variety of test voltages. Although a precision temperature-compensated zener in a constant-current power supply would do the same job—particularly if the zener is held in a constant temperature environment—don't substitute a **simple** zener reference in place of the batteries.

measurement errors

Make a few drift runs on your crystal and audio standard to find out what the errors are. Similarly, make a few runs on the vfo power supply to find out how far it actually drifts. In a complex setup like this, a number of uncertainties exist, including vfo supply voltage, crystal standard error, interpolation error in reading the audio generator, audio generator drift and frequency multiplication error. In the work described here, every effort was made to maintain cumulative measuring error within 10 Hz.

cause of the rapid warmup, I suspect that much of the warmup drift is caused by C2. Replacing this capacitor with a fixed air padder might help—albeit slightly. A larger, high-Q coil might also help.

Although I used carbon composition resistors in all the vfo models that I built, precision film resistors chosen for temperature coefficient may make some slight improvements in operating-point stability. However, these improvements would all be small, and I suspect the cost of the unit would increase considerably before there would be

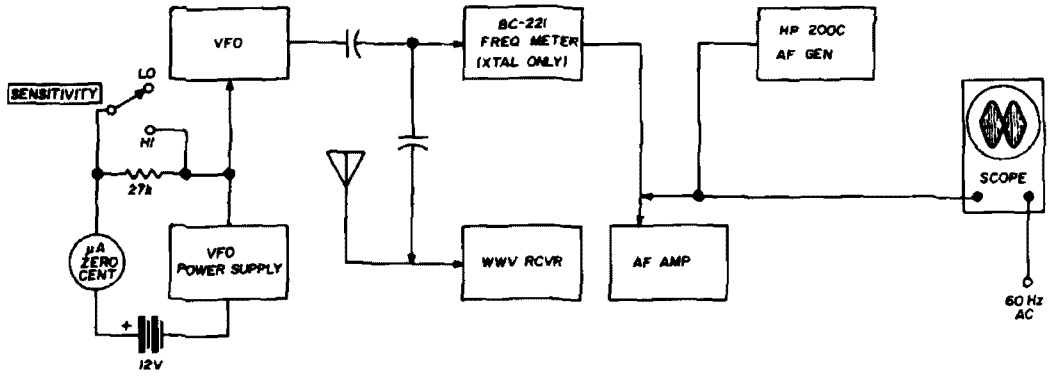


fig. 10. Test setup for measuring vfo drift.

The last source of error is usually neglected in vfo descriptions—environment. My basement is relatively cold and drafty with a fair amount of temperature variation. A number of runs were repeated when data was suspected due to temperature variations. My "standard" in this case is a photographer's darkroom thermometer. With the exception of drift run number 1, the drift curves are those which had the least ambient temperature variation during the run. Furthermore, no drift run was started less than 24 hours after any previous run, or after I did any soldering on the circuit.

summary

It may be possible that with further experimentation, the stability of this circuit could be improved. However, with the instrumentation I am using, it would be difficult to evaluate any improvement accurately. Because of the rapid warmup, I suspect that much of the warmup drift is caused by C2. Replacing this capacitor with a fixed air padder might help—albeit slightly. A larger, high-Q coil might also help.

Although I used carbon composition resistors in all the vfo models that I built, precision film resistors chosen for temperature coefficient may make some slight improvements in operating-point stability. However, these improvements would all be small, and I suspect the cost of the unit would increase considerably before there would be any substantial improvement in drift characteristics.

In conclusion I would like to thank Captain Lee, W3JM (ex-W3JHR), for this remarkable circuit; few solid-state circuits are as bug-free and tolerant of component values as this one. The transistors I have used in the course of my experiments have run the gamut—germanium and silicon, pnp and npn, pullouts, unmarked surplus, high and low beta, hf and vhf types have all been used interchangeably with no changes except polarity of the power supply. All produced remarkably similar—and good—results.

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1. P. H. Lee, W3JHR, "A Stable Transistorized VFO," CQ, September, 1963, p. 25.
2. J. R. Fisk, W1DTY, "Stable Transistor VFO's," *ham radio*, June, 1968, p. 14.
3. E. O. Seiler, W8PK, "A Low-C Electron-Coupled Oscillator," *QST*, November, 1941, p. 26.

ham radio

a plug-in integrated-circuit frequency calibrator

You can use your existing receiver calibration crystal and tube heater supply for this IC marker generator

E. H. Conklin, K6KA, Box 1, LaCanada, California 91011

The new sub-bands reserved for amateurs who have taken advantage of the incentive licensing laws are now a fact of life. Operating in these bands is a real privilege for those who have improved their operating skills and technical knowledge.

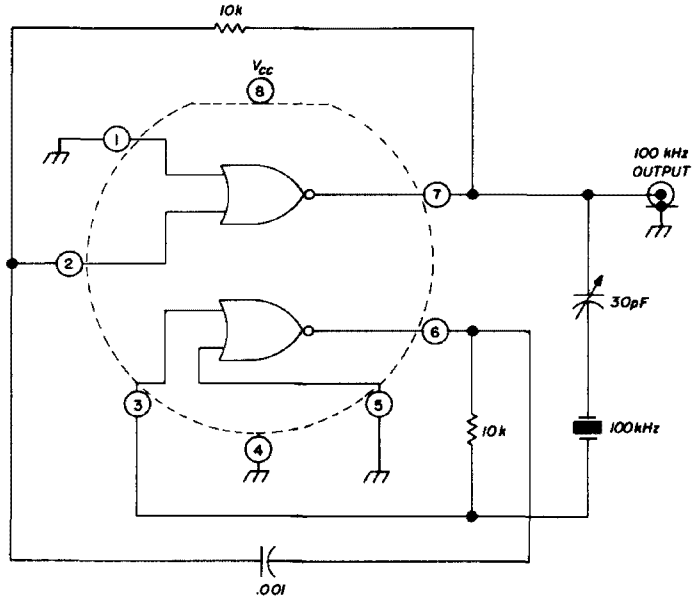
If you don't have your advanced or extra-class ticket yet, you're probably going to need something better than the frequency calibrators built into most amateur receivers. Instead of having to worry about only two frequencies (at the old band edges), you must now be able to determine accurately the edges of the new sub-bands. Even if you're in the top 5000 who are privileged to operate down in the "basement," the little solid-state plug-in calibrator described in this article is just the answer to "where am I?"

Those who may have missed the insert in **ham radio** showing the new frequency allocations can get a copy by writing to the editor. You might like to thumbtack this handy operating aid near your receiver or transceiver as a reminder.

Presented here is a simple plug-in frequency calibrator using the crystal in your

existing calibrator. No extra power is needed. You need only the power used by your present calibrator tube heater. Most calibrators put out markers at 100-kHz intervals. If you have a dial with 1-kHz accuracy, such as the Collins S-Line or Heath SB-300, an output of 10 or 5 kHz is useful. You can build this calibrator for about one dollar above the cost of the integrated circuits. Frequency division will be limited only by the number (within reason) of JK flip-flops you're willing to include.

fig. 1. 100-kHz crystal oscillator circuit using a Fairchild μL914 dual-gate integrated circuit. Grounded pins 1 and 5 lead to unused inputs.



analysis

Switchable heater current appears at the cathode pin of the socket in equipment such as the Swan 500, Heath SB-300, and Collins 75S-3. In the Swan 500, you'll need a larger filter resistor for the plug-in calibrator because of the 12-volt heater supply. If you're using a Heath SB-300 or Collins 75S-3, you must place a jumper across the respective cathode resistors (R17 in the SB-300 and R45 in the 75S-3) to make use of the switch in that line. This power can be rectified and filtered. With low-current drain IC's such as the Fairchild 9093 JK flip-flop, a half-wave rectifier is sufficient; with more drain, a bridge circuit is advisable. Blocking capacitors will be needed in the rf ground-return and output leads to isolate the new power supply negative and the former tube's plate supply.

the circuit

A simple crystal oscillator using two IC gates as the transistors in a multivibrator circuit requires only two 10k resistors and one 0.001 μF capacitor. It will operate from 100 kHz to 1 MHz with no change. The dual-gate Fairchild μL914 performed well in the oscillator circuit[1,2] of **fig. 1**, as did the Motorola MC-724P quad gate. The latter is shown in **fig. 2**. The unused gate inputs are grounded to cut off unnecessary collector current. However, they can be used for isolation amplifiers, R-S flip-flops and other circuits. The cost is around $1.08.

Oscillator output is capable of driving a JK flip-flop as a frequency divider without any intervening circuit. A single flip-flop will divide by two; a dual in-line plastic Motorola MC-790P provides division by four at a cost of only $2.00. Two of these can provide decade division with an output every 10 kHz from a 100-kHz calibrator crystal.

Many other types that draw less current are available, but this is of little advantage except in power supply filtering requirements.* The 5-volt Motorola MC-838P decade divider will produce 50-kHz and 10-kHz harmonics while drawing less current than the 3.6-volt MC-790P. It can be "blocked" into putting out 25 kHz increments instead of decades if desired.[3]

* WØKPZ has pointed out (**QST**, February, 1968, p. 55) that the Motorola MC-778P has less than one-quarter the current requirement of the MC-790P. The MC-778P is in the milliwatt MRTL series, costs only 35c more, is a type-D flip-flop, has lower but adequate output and doesn't require as rapid an input-pulse fall time as the MC-790P to make it toggle. Unless the power supply is filtered adequately, the MC-790P may count power-supply ripple, causing a rough note or incorrect frequency division. **Editor**

construction

I used Vector board with alternate 0.1-inch hole spacing. I drilled a few additional holes to accept the fourteen pins of the dual in-line IC's. Two other types of Vector board that accept the IC's without any drilling are now available. A piece about 2½ x 3½ inches fits nicely into the S-line receiver. It is slotted in one corner to bridge across the stub of the Vector 7-pin tube-base plug. The leads are wired to seven convenient pins, not only to provide connections, but also to attach the plug securely to the board.

With a half-wave rectifier, using a 50-volt "glass amp" diode and two small Japanese 100-μF, 10-volt capacitors from Poly Paks, a filtering and dropping resistor of 82 ohms is required for two IC's using a 6.3-volt supply. This will vary with IC types and the voltage source, so a larger resistor should be tried first to be safe.

The crystal wasn't furnished with a trimming network of zero and negative temperature coefficient capacitors. This circuit is possible here and is advisable. A JFD VC28C piston trimmer was used, providing multiturn adjustment. A larger trimmer could have been used.

The 0.001-μF capacitor (0.1 also worked) and ½-watt, 10k resistors complete the mounting. I used Belden 8430 phono pickup-arm cable, separated into single strands, for wiring the IC's.

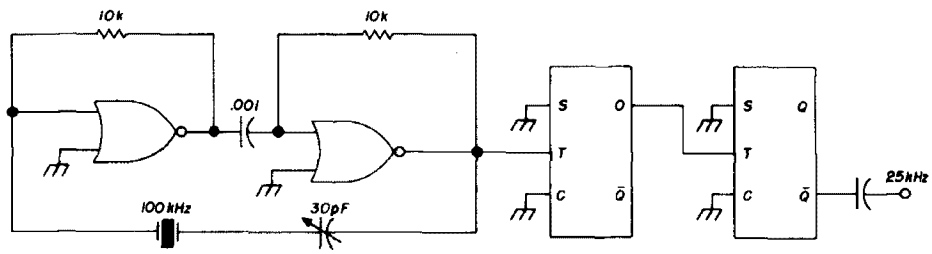
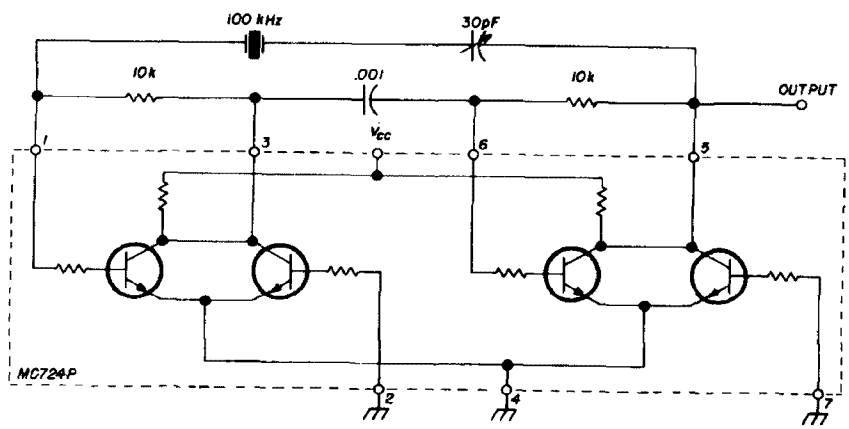


fig. 3. System using two gates and two JK flip-flops to provide 25-kHz markers with a 100-kHz input.

testing

Oscillation can be detected with a vacuum-tube voltmeter, even without an rf probe. Dc will be on the various leads, averaging about half the supply voltage. Care must be taken that there are no short circuits between IC leads, that all connections are conducting, and that all are in their correct places. Most crystals will os-

fig. 2. 100-kHz crystal oscillator circuit using a Motorla MC724P quad gate IC.



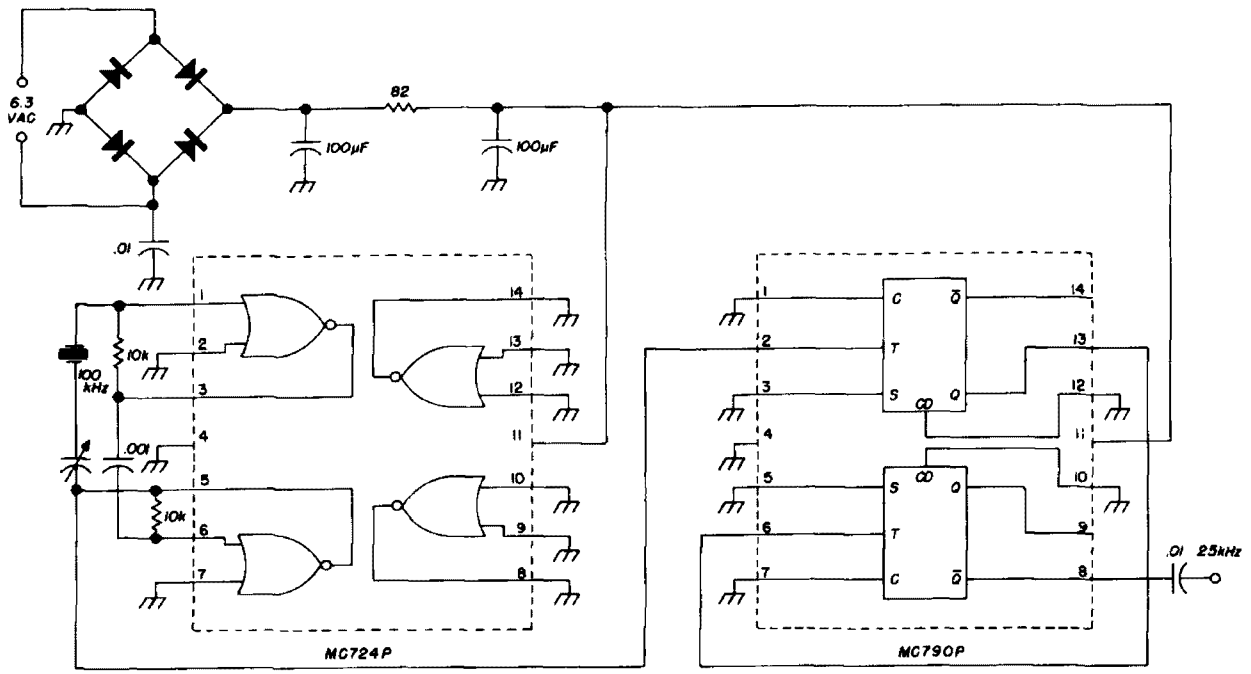


fig. 4. Complete circuit of a simple frequency calibrator that may be plugged into an existing calibrator socket in your receiver. Power supply is derived from the 6.3 Vac filament supply.

cillate in the circuit, but a few take off at a harmonic. A capacitor between 50 and 200 pF across resistor R2, which is in series with the 0.001 μF capacitor, is said to correct this.[1]

Both VK3TE and I found the 100-kHz points from an MC-790P divider to be weaker than intervening points. No correction is offered as yet for this, except to use the Fairchild 9093.

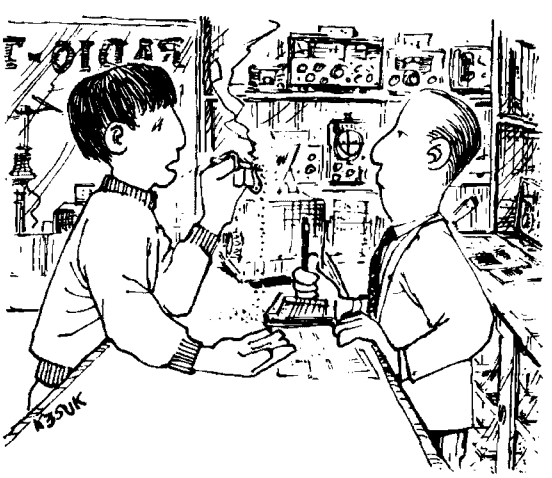
operation

When the power switch is turned to "calibrate," the ground return to the 6.3 volts is applied to the new calibrator board, which replaces the calibrator oscillator tube. The output passes through the original small antenna-coupling capacitor or "gimmick" to the receiving antenna. The new pips appear on the dial at the appointed places. With 10-kHz output, these are adequate to calibrate the intervening nine 1-kHz points by interpolation. If another flip-flop is added to produce 5-kHz harmonics, all incentive band edges are marked if the crystal is set properly—and recently—on WWV. After November 1969, 25-kHz output alone will mark all incentive band edges because the 14,235-kHz limit will not be in effect.

references

1. R & R Electronics reprint accompanying their μL914 printed-circuit crystal-oscillator kit.
2. "Build an IC Crystal Calibrator," *Radio-Electronics*, February, 1968, p. 32.
3. E. H. Conklin, "Calibrators and Counters," *ham radio*, November, 1968.

ham radio



"A slug-tuned coil and a silvor-mica capacitor please . . . and would yuh tune them for 50.42 MHz?"

the real meaning of

noise figure

Jim Kennedy, K6MIO, 5465 East Hampton, Fresno, California 93727

The availability of excellent solid-state devices makes it possible for just about any serious vhf or uhf experimenter to build a receiver with a noise figure that was unheard of only a few years ago. Today's transistors, even in the 400- and 1200-MHz region, are competing with uncooled parametric amplifiers.

Although necessary, a low noise figure in the receiver front end doesn't necessarily guarantee optimum performance of a receiving system. The receiver is only one element of the total receiving system. The performance of the antenna and transmission line can have, in the presence of a low noise receiver, just as important an influence on over-all receiving performance as the receiver itself. The receiver noise figure is a measure of the receiver's performance only. The true measure of the system's ability to detect a signal entering the antenna is the system noise figure (assuming constant pre- and postdetection bandwidths).

There was a time when, due to a poor receiver, the system noise figure was often nearly the same as that of the receiver. As noise figures moved lower, the practice of using the noise figure as a measure of performance became cumbersome. The use of noise temperature is now more meaningful and convenient. Of course, noise figure and noise temperature are related, since they both purport to measure the same thing. In reality, however, the noise temperature is the more basic unit and is actually easier to deal with, both in understanding concepts and making practical calculations.

I've used some mathematics to explain relationships and fundamental concepts. The math is quite straightforward, but it's easy to get sidetracked when there are several terms with nearly the same definitions. The list of symbols and definitions in **table 1** should help you to find your way through the mathematical expressions.

thermal noise

To establish a common basis for the discussion, it seems appropriate to begin with the meaning of the terms noise temperature and noise figure. Noise figures are actually defined in terms of noise temperature; consequently noise temperature needs to be explained first, but not without providing some background.

Any substance that is warm (and by "warm" I mean above absolute zero) contains some electrons that are free to move about in that substance. The amount of energy contained by these electrons increases as the temperature increases, and an increase in energy means an increase in the average speeds of the free electrons. However, moving electrons constitute an electric current. Since we can, in general, assign a resistance to a particular piece of that substance, there should be a voltage developed across the substance due to these thermal electron currents since $V = IR$.

The direction and speed of any given electron is random and fluctuates wildly as the electron moves through the substance and collides with the relatively immobile atoms (ions) that make up the bulk of the substance. The final effect is that, in most substances, there is no net current in any direction on a long-term average, but rather a series of random positive and negative pulses. In other words, the voltages developed by these currents constitute thermal agitation noise.

Since the currents increase with temperature, the noise power likewise increases with temperature. Further, as the pulses are random, they are spread out over a very broad range of frequencies. It develops that, if we look at the power contained in a given passband, the value of that power is independent of the actual center frequency of the passband. Specifically, Nyquist's formula states:

$$p = kTB$$

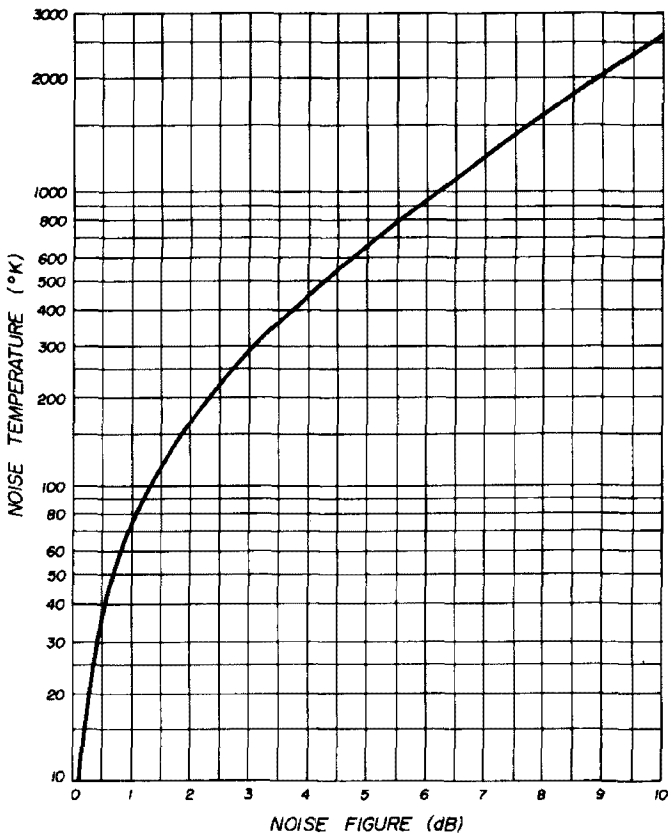


fig. 1. Noise temperature vs noise figure.

table 1. Symbols used in the discussion end mathematical description of noise.

B	**bandwidth**
F	**noise figure**
f	**noise factor**
G	**stage gain**
I	**thermal current**
L	**transmission line loss factor**
P	**thermal noise power**
R	**resistance**
T	**absolute temperature (°K)**
T_A	**antenna noise temperature**
T_{L}	**total transmission line noise temperature**
T_l	**transmission line noise temperature**
T_P	**transmission line physical temperature**
T_R	**receiver basic noise temperature**
T_r	**receiver noise temperature modified by TL**
T_S	**system noise temperature**
V	**voltage (product of IR)**

where k is Boltzmann's constant, T is the absolute temperature (degrees Kelvin), and B is the bandwidth (Hz). Note that the power is directly proportional to temperature, and at 0^0 K there is no noise.

These facts about thermal agitation enter into the discussion of system performance from two aspects. One is that this mechanism can be a source of unwanted noise and the other, which will now be considered, is that this effect can be used to provide a convenient standard noise source for measuring receiver or system performance. It is in this latter connection that the concept of noise temperature has evolved.

noise temperature

Suppose you have a receiver front end that needs to be evaluated. Further, suppose that its input impedance is 50 ohms. One way to measure performance would be, in principle at least, to terminate the receiver's input with a warm 50-ohm load, attempt to detect the load's thermal agitation noise, and compare it with the noise generated by the front end itself.

To be even more specific, suppose we

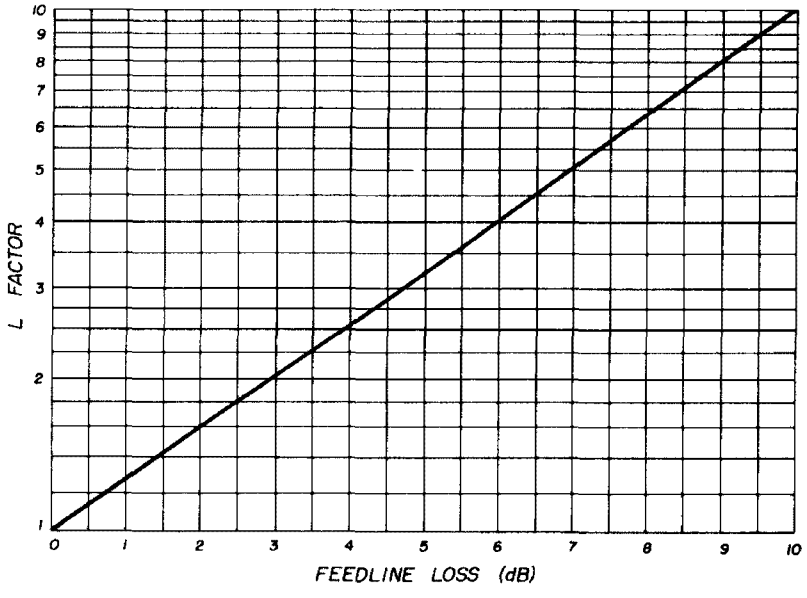
could reduce the resistor's temperature to absolute zero and note the noise level on a perfectly linear S meter. This noise level would be due entirely to the receiver, because there is no thermal noise in the resistor at absolute zero. Now suppose we increase the resistor's temperature until the S meter shows a 3-dB increase. That is, until the resistor thermal agitation noise is equal to that generated by the receiver. We've obviously measured the noise generated in the receiver, but what is it?

We could determine the power in watts from Nyquist's formula, if we knew the exact bandwidth of the receiver, since we already know the resistor's temperature. But a number that depends on the receiver's bandwidth would be somewhat cumbersome for comparing two different receivers. On the other hand, a very convenient and accessible number would be the resistor temperature. For example, if heating the resistor to 400° K would produce the same as that generated by the receiver, we could say that the receiver noise temperature was 400° K. This number is independent of the receiver bandwidth, and we can use it to compare receivers of different bandwidths.

it should be noted parenthetically that this number is meaningful, because the resistor noise and the receiver front-end noise are both random and relatively broadband. In reality the noise the receiver generates may or may not be caused by thermal effects, but we can always compare the amount of noise per unit bandwidth with that which would be produced by a resistor of some certain temperature.

One example of the convenience of using noise temperature as a performance measure is that it is directly proportional to noise power. Hence if, for instance, you were to halve the noise temperature of a system, you would double the signal-to-noise ratio of any signal present. The same can't be said for noise figure.

fig. 2. Curve of factor L vs feedline loss.



noise figure

Before we can define noise figure we must take one intermediate step. Suppose we define a noise factor

$$f = \frac{T}{290} + 1$$

where T is the noise temperature, and the 290 represents the value of room temperature in Kelvin units. The noise factor is a number that varies from 1 to infinity as T varies from zero to infinity.

Now we can define the noise figure as the noise factor expressed in decibels; that is

$$F = 10 \log_{10} f$$

Note that when T = 290° K, room temperature, f = 2, and the noise figure, F = 3 dB. In other words, both f and F are ways of comparing T and 290° K. Note also that the noise figure is in a sense two generations removed from our original unit, noise temperature. A little scratch paper calculation will show that

$$T = (f\text{-}1)\ 290$$

Unlike T, the percentage change in f necessary to produce a given improvement in signal-to-noise ratio depends on the original value of f. This effect is particularly noticeable when f is small, which is precisely the region we wish to consider. Consequently, such calculations, while not too difficult, are not instantaneous. Hence the significance of f and F is more obscure than that of T.

The foregoing has, I hope, established the relationship between T and F and provided some initial insight into the reasons for dispensing with noise figure in favor of noise temperature in what follows.

receiving systems

Earlier it was suggested that the over-all performance of a receiving system depends not only on the receiver but also on the transmission line and antenna. Each of these elements makes a contribution to the system noise temperature, and this contribution is directly additive (another convenience of noise temperature over noise figure). It should be emphasized that the intent is to find a system noise temperature. In this context, when considering the antenna, we are only interested in the antenna's noise performance and not its gain.

A great deal has been written about the sources of noise in various rf amplifiers. It's therefore unnecessary to spend much time on this aspect of the subject, except to note that, in vacuum tubes and transistors, the actual source of most of the inherent noise is usually nonthermal in origin. However, in parametric amplifiers, the primary noise source is thermal, and cooling the paramp will actually improve its noise temperature. Usually the noise temperature of a paramp is quite a bit lower than its physical temperature due to the slightly strange way that thermal noise develops across a negative resistance. But it is related to physical temperature, and hence cooling helps. On the other hand, cooling transistors and tubes will turn them off.

Of course, the noise temperature of any good receiver is determined principally by that of the first preamp and almost entirely by the combined effect of the first two preamps. This can be seen from the general formula for receiver noise temperature.

$$T_R = T_1 + \frac{T_2}{G_1} + \left[\frac{T_3}{G_1 G_2} + \frac{T_4}{G_1 G_2 G_3} + \ldots \right]$$

The subscripts refer to amplifier stage number, and G is the stage gain expressed as a power ratio rather than in decibels. Since all the stage noise figures are usually within the same order of magnitude, the growing gain product in the denominators makes any contribution to the noise temperature beyond the second stage very small, and the series can be truncated after the second term.

transmission lines

The transmission line's contribution comes from two effects produced by a common source: line losses.

The first of these two effects is the more obvious. When a signal travels down a lossy line, the signal is attenuated. This reduces the signal-to-noise ratio, which is equivalent to directly increasing the noise temperature of the receiver. This increase can be expressed by introducing a loss factor, L, which varies from one to infinity and is the ratio of power in to power out for the transmission line. The combined effect

of receiver noise temperature and this aspect of the line loss is

$$T_r = LT_R$$

where T_R is the basic noise temperature of the receiver.

The second effect arises from our familiar physical temperature considerations. The fact that losses exist implies that there is a true resistive impedance associated with the line. (Note this is distinct from the characteristic impedance, which is something else entirely.) Since the line is warm, it actually generates noise due to thermal agitation. Hence, the noise temperature is related to the loss factor, L, and the physical temperature of the line, T_p. Specifically,

$$T_L = (L\text{-}1)\, T_p$$

Note that when line losses are low, but receiver noise temperatures are on the order of room temperature, the predominant transmission line contribution to system noise is the first effect. However, when receiver noise is very low the second, or thermal, effect predominates. Though very few amateurs will encounter such problems, this is commonly the situation with maser amplifiers, and in such cases cooling the transmission line, while often impractical, would improve performance.

The first transmission line formula combines the line effect with the receiver noise temperature, but it's often more convenient to combine the two line effects and state the receiver contribution separately. Defining T_L as the total transmission line noise temperature we can write

$$T_L = (T_p + T_R)\, (L\text{-}1)$$

This is, again, only the transmission line contribution, although it does involve receiver temperature as a parameter.

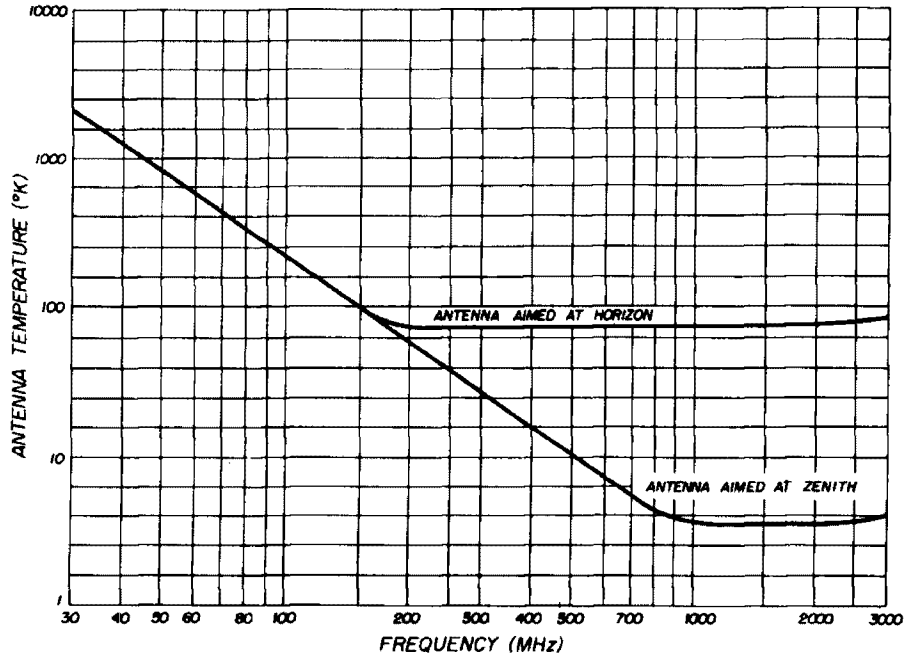


fig. 3. Minimum expected antenna temperature as a function of frequency.*

antennas

Of all the contributions to system noise, antenna temperature is probably the most misunderstood and also the hardest to pin down. First of all, assuming the antenna

* J. D. Kraus, H. C. Ko, "Celestial Radio Radiation," Ohio State University Radio Observatory Report 7, May, 1967; Dicke et al, "Cosmic Black-Body Radiation, *Astrophysical Journal*, volume 142, 1965, pp. 414-419.

to be constructed of good conducting materials, the antenna noise temperature has practically nothing to do with the physical temperature of the antenna. The antenna noise temperature is, rather, a measure of the noise the antenna is actually receiving (as compared to a warm resistor on the end of the transmission line instead of the antenna). The amount of noise an antenna picks up is often, but not always, related to the temperature of the objects the antenna "sees" at uhf, but at vhf the opposite is generally true.

If we consider the uhf case, an antenna on an earthbound circuit sees a large amount of warm earth, and the antenna temperature will be in the neighborhood of 100° to 300° K. This is because the free electrons moving in the warm earth radiate electromagnetic energy in proportion to their temperature.

There is little that can be done to improve the situation of an earthbound circuit, but an antenna that looks at the sky, such as a moonbounce antenna, would have a much lower noise temperature since the sky is much "colder" than the earth.

The effect of the earth is still present, however, since any sidelobe that sees the earth will pick up thermal noise. This effect can be quite serious, and sidelobes are of major concern in many deep-space communications and radio astronomy systems. Careful attention to antenna design with respect to sidelobes can result in antenna temperatures significantly under 50° K, while poor design can result in much higher values (approaching 300° K as a limit). This effect is indicated in **fig. 3.**

system noise temperature

The final measure of the amount of noise power per unit bandwidth that an incoming signal must compete with is given by the system noise temperature. This is just the sum of the receiver, transmission line and antenna noise temperatures,

$$T_S = T_A + T_L + T_R$$

$$T_S = T_A + (T_p + T_R)\,(L\text{-}1) + T_R$$

examples

Having presented this array of formulas, it might be useful to demonstrate them with a few typical examples.

You occasionally hear a heated discussion on the air about whether or not a super low-noise front end is really any better than a mediocre front end on six meters. The crux of such a discussion should center on the influence of sky- and man-made antenna noise on the system noise temperature. Let's join the fray and consider the question.

Antenna temperatures at six meters are quite high. A typical value for T_A would be 3,000° K. The next aspect to consider would be the feedline. In the case of an 80-foot tower, which might have a 90-foot run of RG8/U, the feedline loss, from published data, is about one decibel. From **fig. 2** we see that L is equal to 1.26. Then

$$T_L = (290 + T_R)\,(0.26)$$

Now, to find the system noise temperature all we need do is to try a couple of values for T_R. Suppose a receiver with a good front end has a 1-dB noise figure and a mediocre one, 4 dB. Referring to **fig. 3,** then, using the system temperature formula, the good receiver provides

$$T_S = 3000 + [(290 + 76)\,(0.26)] + 76$$

$$T_S = 3171° \text{ K}$$

On the other hand, the mediocre one provides

$$T_S = 3000 + [(290 + 438)\,(0.26)] + 438$$

$$T_S = 3627° \text{ K}$$

It was obvious from the beginning that the good front end was superior to the mediocre one, but in terms of the over-all system performance, there is less than a 0.6-dB signal-to-noise ratio advantage. The argument will, of course, rage on; but at least now it must be over the relative advantage of a 0.6-dB signal-to-noise ratio improvement.

Let's now shift our attention to a pair of 432-MHz examples. First we'll consider an example paralleling the one above. On 432 MHz, we might find an antenna temperature of 300° K. In this example we'll again assume the feedline loss at one decibel. Now consider the difference between a 2-dB paramp or transistor front end and a 6-dB 416B amplifier. In the former case

$$T_S = 300 + [(290 + 168)\ (0.26)] + 168$$

$$T_S = 587° \text{ K}$$

In the latter case,

$$T_S = 300 + [(290 + 870)\ (0.26)] + 870$$

$$T_S = 1472° \text{ K}$$

Hence the better preamp would make a 4-dB improvement in system noise. The fact that 4 dB is also the difference in the preamps noise figures is purely coincidental. To illustrate that this is so, let's take another quick example of the same sort. Consider a radio telescope with a T_A of 20° K and a feedline loss of 0.1 dB. Now if we use a maser with $T_R = 3°$ K, then,

$$T_S = 31.8° \text{ K}$$

The same system but with a 3-dB front end has

$$T_S = 327.5° \text{ K}$$

Here a 3-dB receiver noise figure improvement has led to more than 10 dB of system improvement.

In the second 432 MHz example, we'll examine the influence of the feedline. Assume $T_A = 300°$ K and $T_R = 168°$ K (2 dB). We've previously shown that when the line loss is one decibel

$$T_S = 587° \text{ K}$$

But suppose we increase the line loss to 4 dB. Then L = 2.55, $T_L = 710°$ K and

$$T_S = 1178° \text{ K}$$

The system degradation is, in this case, 3.1 dB, **not** 3.0 dB. Further, notice that the feedline noise contribution here approaches twice that from all other sources combined.

summary

Of course the foregoing examples are just that: examples. But I hope they have illustrated some of the important aspects of the problem of receiving system performance.

Perhaps it would be well to underscore in a qualitative way some of these points. For a given receiver passband and antenna gain they might be stated as follows.

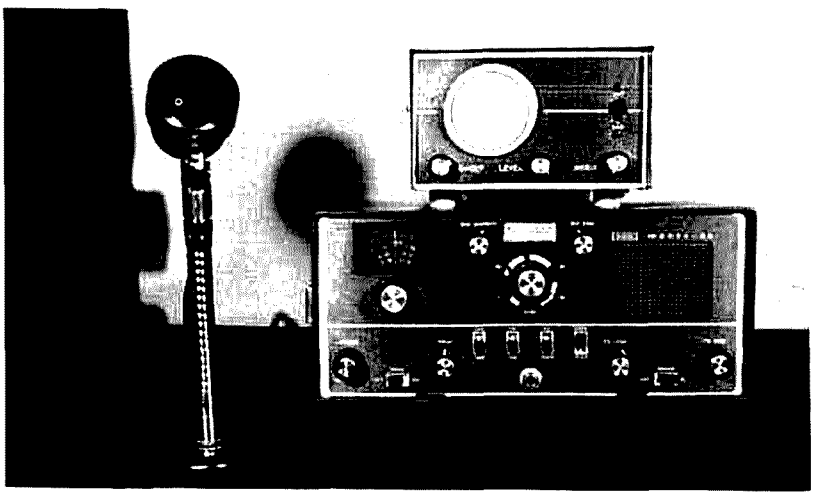
1. Receiver performance is ultimately limited by the signal-to-noise ratio present at the receiver end of the transmission line. (This was obviously the case in the six-meter example.)

2. Assuming the antenna noise temperature is minimized or can't be controlled, the transmission line is a significant factor in establishing this signal-to-noise ratio.

3. Once the signal-to-noise ratio at the receiver input has been maximized, then the noise temperature of the receiver itself can be minimized to produce the least degradation in the input signal-to-noise ratio. The extent to which this is practical depends largely on actual improvement secured, as contrasted by the six-meter example and the maser example.

Finally, it should be clarified that this entire discussion has hinged on noise performance. Obviously, increasing antenna gain will also improve performance. Unless the antenna is looking at a noise source of limited size, such as the sun or a bad power-line transformer, the antenna noise temperature is, in general, independent of the gain. Therefore we could add as a fourth point: put up the biggest, grandest, best antenna you can.

ham radio



a miniature monitorscope

It is pretty hard to beat an oscilloscope like this for less than thirty dollars— it features a surplus CRT

Lee Richey, WA3FIY, R.D. 1, Franklin, Pennsylvania 16323

There is little question of the desirability of a monitor oscilloscope in today's ham shack. The problem with many of us is the availability of the funds necessary to purchase one. The oscilloscope presented here will display both envelope and trapezoidal waveforms when used with a transmitter of 10 watts or more. It uses a very linear internal sweep circuit with variable sweep rates from 30 Hz to 300 Hz, or it can be switched for external horizontal input signals with a full-scale sensitivity of about 0.5 volt rms. The scope also includes a blanking circuit that blanks the trace when no rf is present. If you have a well-stocked junk box, you can construct this scope for just a few dollars. If not, you can still build it for under $30*.

the circuit

The monitorscope is built around a 902A cathode ray tube available from many surplus dealers. The 902A is ideal for this application for several reasons. First, it is available for less than $3 in many areas of the country. Second, it has an accelerating voltage of 400-600 volts, easily obtained with low-cost components. Third,

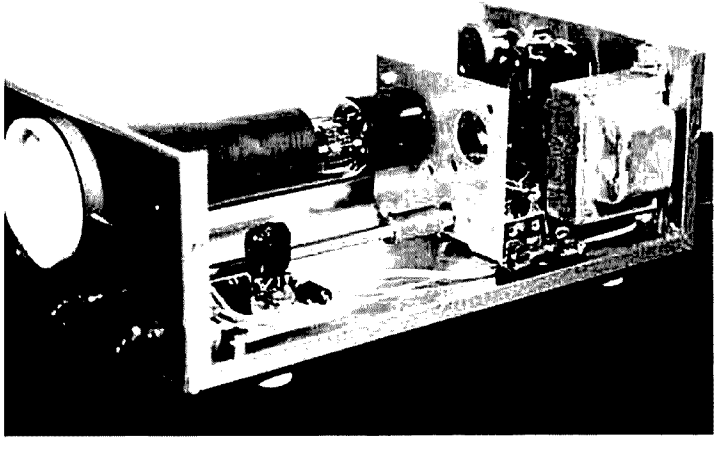
the deflection sensitivity is approximately 100 volts per inch, allowing the use of transistorized deflection amplifiers. Fourth, the tube is small—2 inches in diameter, 7 inches long—and uses an octal socket.

The power supply is a voltage tripler using silicon diodes and a low-voltage power transformer. It produces 175 volts dc for the horizontal deflection amplifier and associated circuits and 525 volts dc accelerating voltage.

The horizontal amplifier consists of a source follower junction FET input stage for high impedance input and a voltage amplifier using a low-cost, high-voltage silicon transistor. The input impedance of the FET stage is 500k ohms, and the voltage gain of the amplifier is approximately 100. The undistorted output is over 150 volts peak-to-peak and gives more than adequate sweep width. Maximum sweep width is about 1½ inches with the horizontal amplifier supply voltage of 175 volts.

For normal waveform monitoring this is sufficient, but if desired, it may be increased to full scale by using a voltage doubler power supply producing approximately 550 and 275 volts dc. A higher-voltage horizontal output transistor, such as the RCA 40264, would also be necessary.

Internal layout of the monitorscope. The horizontal sweep circuit board is in the foreground. The power transformer is mounted on the rear wall.



The internal sawtooth generator is basically a unijunction transistor relaxation oscillator. The sawtooth voltage across the timing capacitor is linearized by a constant-current source (transistor Q1 on the schematic) charging the capacitor. From the equation:

$$E = \frac{TI}{C}$$

Where: E = voltage across the capacitor
I = charging current
C = capacitance
T = time

It can be seen that if the charging current and the capacitance are held constant, the voltage across the capacitor is directly related to time. This approach works very nicely in practice, as well as in theory, and yields a very linear sawtooth voltage.

The current, while held constant for any given sweep frequency, is made adjustable to set the sweep frequency. The supply voltage for the sawtooth generator is regulated at 11 volts by reverse biasing the emitter-base junction of a Fairchild 2N5135 silicon transistor. This "diode" costs $.25 compared with over $1.00 for a typical zener of the same rating.

At this point it might be worth mentioning that many of the low-cost, general-purpose silicon transistors can be used as zener diodes by reverse biasing their emitter-base junction. To tabulate a few:

2N5135	– 11 volts
MPS3704	– 9 volts
MPS706	– 8 volts
2N5138	– 7 volts
2N5134	– 6 volts

The zener voltages listed were measured on twenty or thirty devices and represent an average.

* The semiconductor package is available from the author for $6.00 plus tax and shipping. The total package is also available, including Minibox, CRT, transformer, semiconductors, passive components, line cord, and all hardware, switches, and knobs. Price: $28.75 plus tax and shipping.

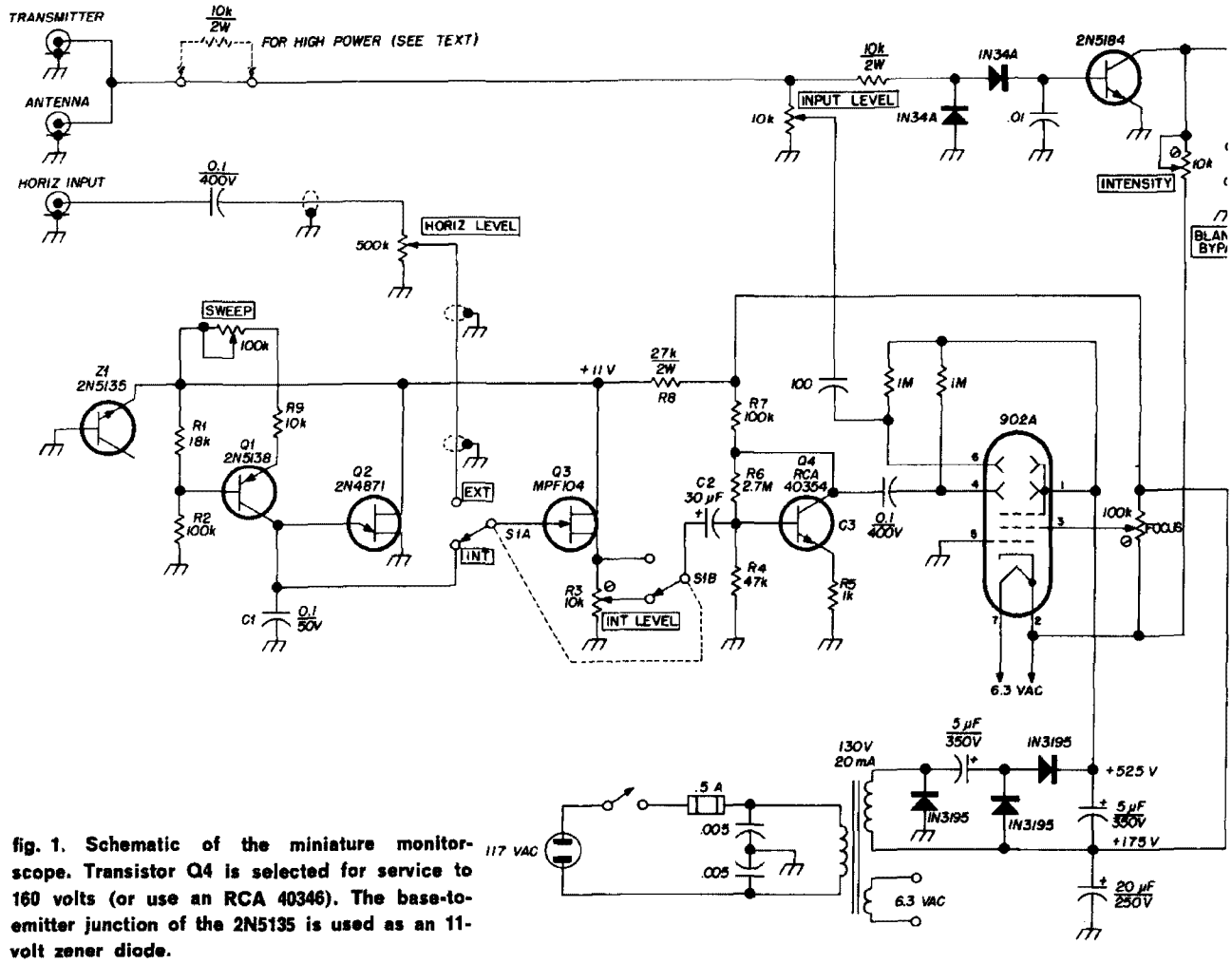


fig. 1. Schematic of the miniature monitorscope. Transistor Q4 is selected for service to 160 volts (or use an RCA 40346). The base-to-emitter junction of the 2N5135 is used as an 11-volt zener diode.

The blanker circuit is composed of a transistor switch, Q5, connected in series with the intensity control and the cathode of the CRT. A diode rectifier circuit supplies base current to the switching transistor when rf is present at the input terminals, allowing cathode current to flow and making the trace appear. However, with no rf present, the transistor is cut off, and the trace is blanked. This feature prevents "burning" the CRT and adds to its useful life. For special applications the blanking circuit can be bypassed by closing switch S2.

The vertical plates of the CRT are capacitively coupled to the input jack through the level-control potentiometer. This type of connection requires no tuned circuits and simplifies operation of the scope. This input circuit works well with coax feedline, but balanced feedlines will require either a tuned circuit transformer or a pick-up loop placed near the transmitter tank circuit or the antenna tuner. In any case, the scope requires about 50 volts peak to peak for usable deflection.

construction

The monitorscope is mounted in a 10 x 6 x 3½-inch Minibox. A combination of circuit board and point-to-point wiring was used. While the layout and wiring are not critical there are several precautions that should be taken. Since no CRT shield was used, in the interest of low cost, any magnetic fields present in the circuit should be as far away from the CRT as possible. The primary source of such fields, the power transformer, is located behind and as far to the side of the CRT as the chassis will allow. The leads carrying ac power to the

front panel switch were shielded as a further precaution.

Mounting the CRT posed the biggest problem. After several unsuccessful attempts, I finally decided to use a soft foam plastic drinking cup, now almost universally available. First, cut a 2-3/16-inch-diameter hole in the front panel of the enclosure. At this point I would recommend using a pair of tin snips or a nibbling tool and then finishing with a file. Fly cutters and hole saws do a nice job, but they can be quite dangerous when the work piece is not adequately held down.

When the hole is finished, cut the bottom out of the plastic cup, insert the CRT base first through the bottom of the cup, and then insert this assembly through the front panel and pull it up snugly. The base of the CRT is supported by its octal socket mounted on an L-shaped bracket. **Do not** permanently mount the CRT socket until the scope has been tested and the tube rotated to make the trace horizontal; pin 1 of the CRT should be positioned near the top for preliminary alignment.

The excess foam cup extending through the front panel can be trimmed with a razor blade or sharp knife, leaving enough material to act as a bezel. A piece of plastic tape should be wrapped around the cup where it contacts the front panel.

Because they are seldom adjusted once set, the focus and intensity controls are mounted on the rear of the enclosure. Also, the blanker bypass switch S2 is mounted on the rear of the enclosure, because it is only used for special applications.

The leads carrying rf are kept as short as possible to the extent that the input-level potentiometer is mounted on the bracket that holds the CRT socket. A shaft extension is used to bring the control to the front panel. The blanking circuit is wired on a terminal strip and is mounted directly behind the level control.

Power supply components are wired to a terminal strip, which is mounted to the rear wall of the enclosure. The horizontal amplifier and sawtooth generator are built on a 2 x 2-inch piece of G-10 epoxy glass Vector board with holes spaced on 1/10-inch centers. The components are mounted on one side of the board with their leads extending through to the other side. The leads are bent over and soldered to the leads of the other components to which they attach. Short pieces of bare wire are used for supply leads and grounds. This is a very quick and easy way to build "printed" circuits without the fuss and mess of screens, resistors, etchants, drills, etc. Standard precautions should be taken with the semi-conductor devices.

When complete, the circuit board is mounted on ¼-inch spacers using number-4 hardware in the area behind the horizontal-level control and beside the CRT. Lead dress to the board is not critical, but shielded wire is recommended for the FET input circuit.

testing, adjustment and use

After completing the construction and double checking your work, apply power to the scope and check the three supply voltages (11 volts dc, 175 volts dc, 525 volts dc). If they are within reason, close S2 and advance the intensity control until the trace becomes visible. At that point, adjust the focus control for the sharpest possible line or dot. Switch to internal sweep, and advance the internal level control to the point where the trace no longer lengthens or where the trace is approximately 1½

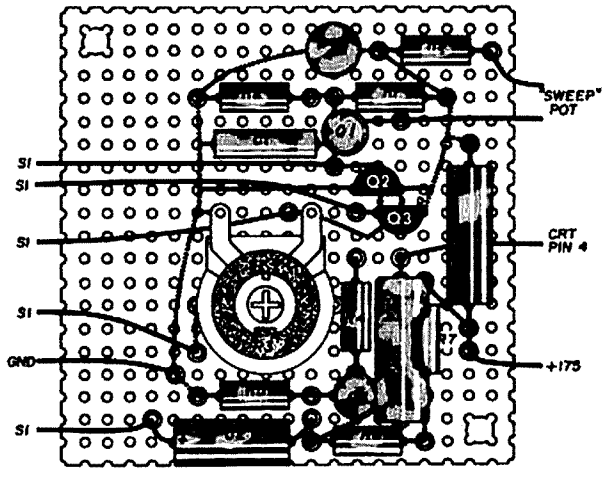


fig. 2. Circuit-board layout for the horizontal amplifier.

inches long. For best performance the 2.7-megohm bias resistor for the 40354 (Q4) should be adjusted so that the collector voltage is one-half the supply voltage (87 volts in my case). Rotate the CRT slightly to bring the trace horizontal. You may want to mark the mounting bracket and permanently mount the CRT socket at this point.

The monitorscope is now ready for use. Connect a piece of coax between the scope and the transmitter and connect the antenna to the scope. For envelope patterns use the internal sawtooth sweep, adjust the input-level control for on-scale displays, and adjust the sweep control for low sweep frequency. Open S2, and the trace will only appear when transmitting.

For trapezoidal or "bow-tie" patterns, connect the horizontal input to a point in your transmitter that has 0.5 to 2 volts of audio, and switch the scope to external horizontal input. With maximum audio applied to the transmitter, adjust the horizontal level to slightly less than with the internal sweep. Trapezoidal patterns are quite useful for monitoring the modulation percentage of a-m transmitters. "Bow-tie" patterns are obtained from double sideband, suppressed-carrier transmitters and are useful for checking amplifier linearity.

Single sideband transmitters, however, will require the use of the envelope pattern. Checking the performance of an ssb transmitter generally requires the use of a two-tone audio generator, which could be built into the monitorscope if desired. For waveform analysis and other information, consult the **Radio Amateurs Handbook**, the **Radio Handbook,** or **Radio Communications Handbook.**

When this scope is used with transmitters of 400 watts **output** or more, insert a 10k ohm 2-watt composition resistor between the input jack and the level control (the point marked X in the schematic). Also, for use with high-power rigs, SO-239 or similar types of rf connectors should be used in place of the phono jacks. There are actually several ways to connect rf to the monitorscope. **Fig.** 1 shows the method just described, which is the way it's used at my station.

Some transmitters, such as the Heath HX-10, have a monitor jack provided for connecting an oscilloscope or other devices. Usually a small-value variable capacitor is connected to the transmitter rf output jack.

As mentioned before, the CRT is unshielded, so use care in the placement of this scope when in use. If it's located too close to power transformers or rf fields, distortion of the trace will occur. If it's impractical or impossible to keep the scope out of such fields, I would recommend using a CRT shield such as Millen 80802-A*.

One final note in regard to centering the trace. As you will notice, no centering con-

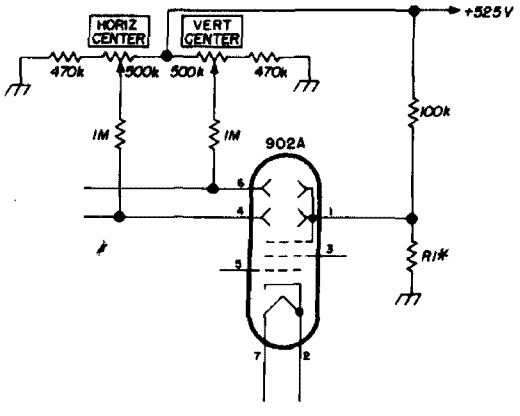


fig. 3. Circuit modification for trace centering. Resistor R1 is adjusted for approximately 400 volts on pin 1 of the 902A; start with 470k ohms.

trols have been used on this scope, as the 902A is fixed centered. While the trace is not perfectly centered, it is good enough for all but the most particular individuals. These controls were left off for simplicity and lower cost, but if you desire to move the trace around a bit, **fig. 3** shows a few modifications that can be made. The resistor marked with an asterisk should be chosen to develop about 400 volts accelerating potential (pin 1) at the CRT with the focus and intensity controls adjusted for use. Start with about 470k ohms and adjust accordingly.

* Available from Millen distributors for $7.30.

ham radio

seven-tenths-wavelength ground plane for two meters

Ted Capsanes, W3WZA, 6747 Riverdale Road, Riverdale, Maryland 20840

The subject of gain antennas will often produce a varied response when taken under discussion by amateurs. This is especially true in vhf mobile applications where the system uses a grounded vertical as the radiating element.

The purpose of this article is to point out some of the basic characteristics of the ground plane, and to show how these properties were used to adapt an existing antenna to 147 MHz.

characteristics

Commercial mobile antennas designed for the frequency spectrum of 140-170 MHz have certain structural characteristics that distinguish them from their lower-band counterparts. At these higher frequencies, the heavy-duty spring mount base is not practical. Gain antennas within this range are generally limited to half-wave or five-eighths-wave radiating elements that are base loaded to extend the electrical length to three-quarter wavelength. The matching network is sealed within a thin cylinder that forms the mounting base for the entire antenna.

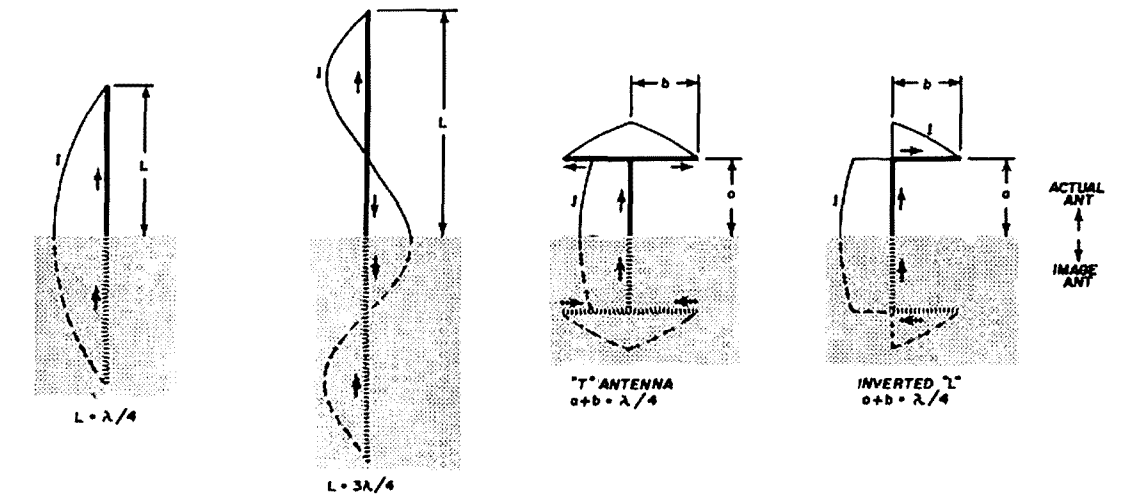
The electrical properties of a grounded vertical are shown in **fig. 1** and **2.** The common quarter wave is widely accepted because the feed point is not reactive and presents an impedance of suitable value. The lobe pattern of a half-wave antenna provides some gain, but the element end presents a current null that can't be matched directly to common transmission lines. A three-quarter-wavelength radiator is an ideal match for 50-ohm transmission line, but low-angle lobes are secondary. With more energy being directed toward the higher angles, direct-wave field strength, at zero elevation angle, is less than that of the quarter-wave antenna.

Most vhf amateurs recognize the five-eighths-wavelength antenna as the ultimate in a single-element radiator, but many don't know the true gain of the device. This is sometimes brought about by misleading literature that uses the dipole or isotrope as the reference source. The top curve of **fig. 2** is convenient for determining true zero elevation angle gain under ideal conditions (perfectly conducting ground).

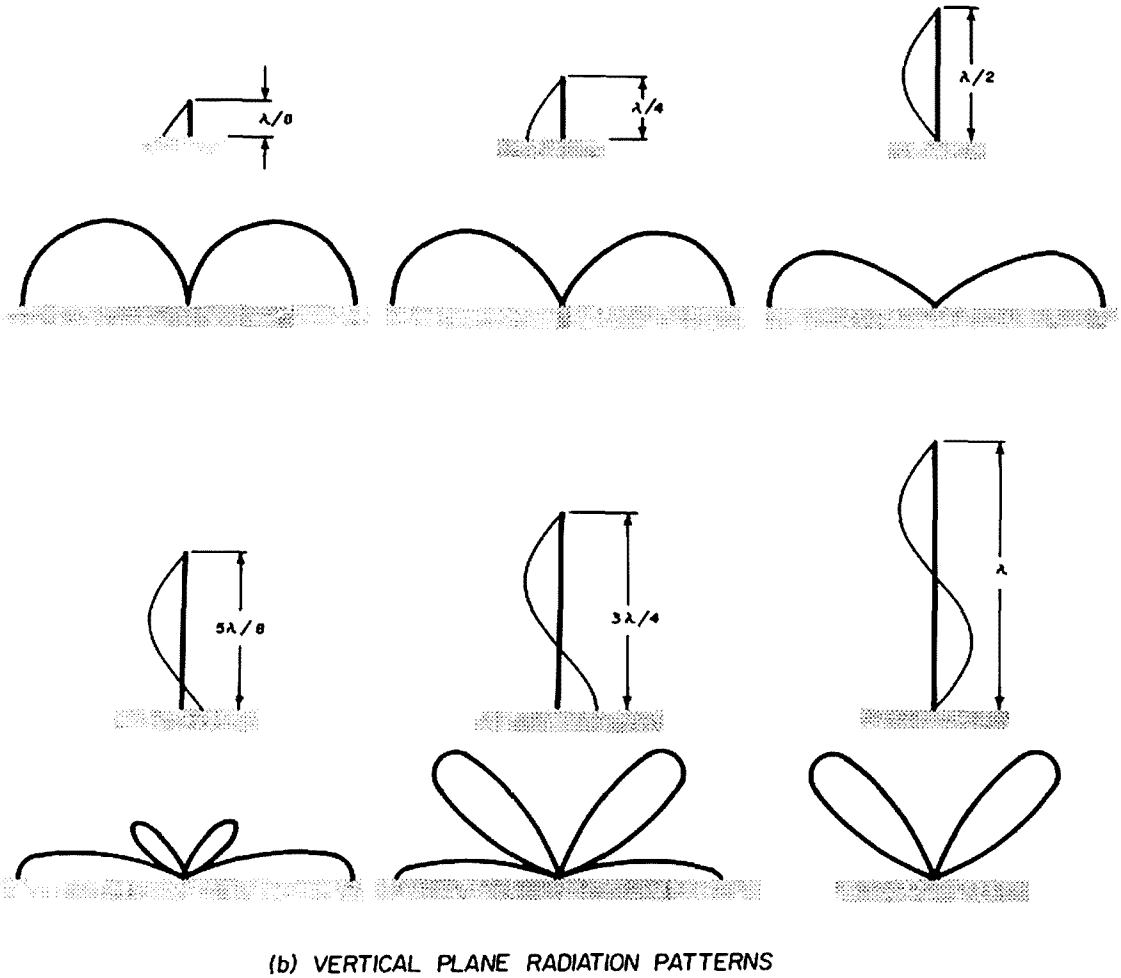
construction

These facts presented an interesting challenge. I wanted to evaluate some Motorola equipment in the fm portion of the two-meter band. However, the only permanent antenna system on the car was a quarter-wave six-meter whip, using a 232-Series Master Mobile tapered spring base. I couldn't see any way to utilize this system as a gain antenna in any conventional manner.

The only practical place to locate a matching network was at the coax terminal point inside the trunk of the car. As stated before, the purpose of the loading network is to extend the electrical length of the system to three-quarter wavelength. This means that the junction point of the



(a) CURRENT DISTRIBUTION



(b) VERTICAL PLANE RADIATION PATTERNS

fig. 1. Characteristics of grounded vertical antennas.

loading coil and the radiating element will exhibit a high impedance with a half-wave radiator, and the impedance will decrease as the radiator length is extended.

Seven-tenths wavelength appeared to be a good compromise. The zero elevation

angle gain was better than that of a quarter-wave whip, and the impedance at the junction point approached minimum. The over-all length of the antenna (including base) is 52-1/2 inches for 147 MHz. The

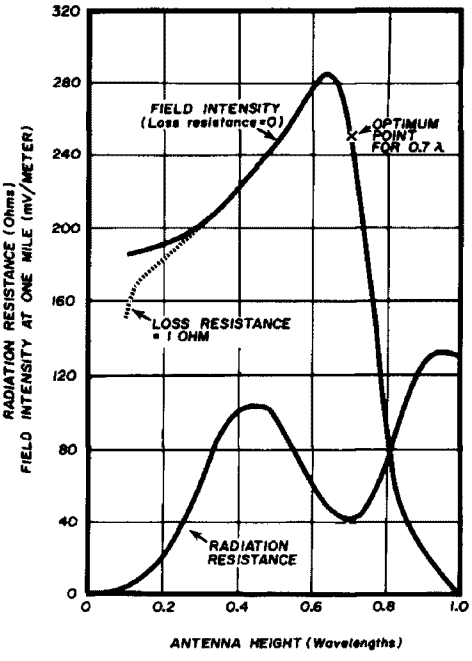


fig. 2. Radiation resistance and field intensity as functions of antanna height.

gain slope in this area is very steep (**fig. 2**), and a difference of one or two inches will alter the characteristics considerably.

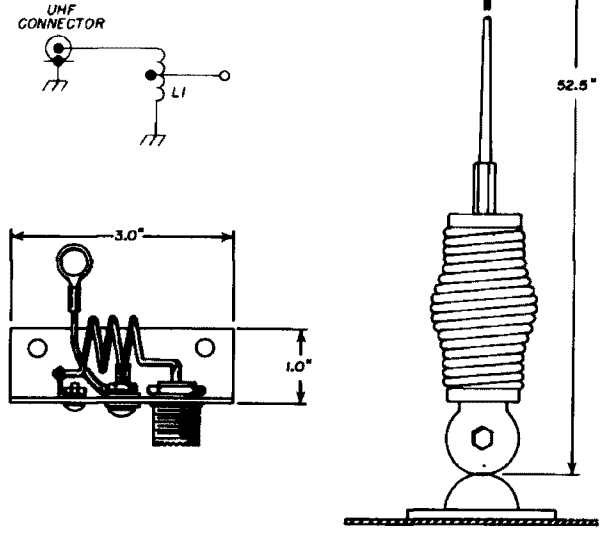
Fig. 3 shows the electrical and mechanical characteristics of the matching network. The device is built on a three-inch length of aluminum angle, with holes spaced to accommodate two of the mounting screws of the existing base. The coil is wound around a ½-inch form using number-14 bus wire. A teflon feedthrough, with a 6-32 screw, serves as the junction point for the coil tap and the antenna base coupling wire. This wire should be cut so the terminal lug center-to-center dimension is 1-3/4 inches. Typical solder lugs provide sturdy mechanical support of the coil at the tap point and ground end.

Electrically, L_1 supplies the requirement of two coils. The 1¼ turn between the uhf connector and the antenna is the series loading inductance. The remaining portion provides the shunt impedance required to cancel approximately 72.5 ohms of capacitive reactance.

Part of the capacitive component of this reactance is due to the mechanical structure of the Master mount and its relationship to the automobile body, when properly installed. This small amount of X_c can be overlooked at the lower frequencies but must be compensated for at vhf. Some additional strays are introduced with the installation of the matching network.

For a period of time, the development of the antenna overshadowed the original task of equipment evaluation. Practical

fig. 3. Mechanical and electrical details of the antenna. L1 is 4 turns no. 14 bus wire, 3/4" long, 1/2" diameter, tapped 2-3/4 turns from ground end (total inductance = 0.16 μH).



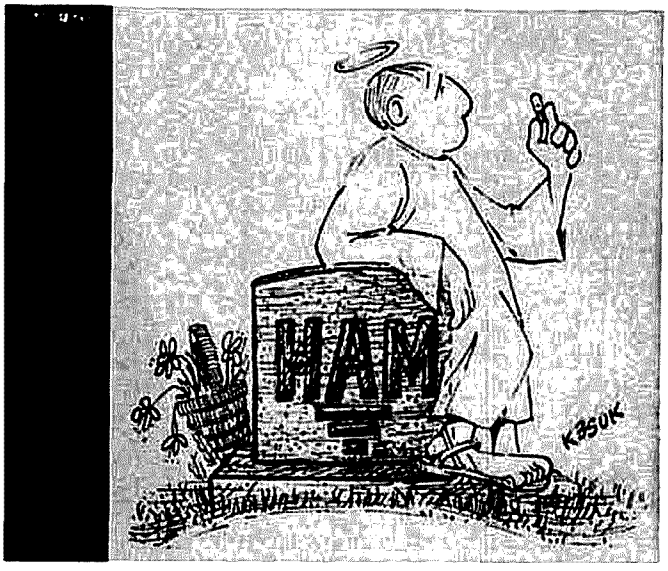
problems had to be discovered and corrected by trial and error. A seven-tenths-wavelength antenna is not conventional, but it does have useful applications. The purpose in pursuing the project was to prove that it could be done.

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J. D. Kraus, "Antennas," McGraw-Hill, New York, 1950, pp. 315, 316.

ham radio



"Just 15 lousy little milliamps"

safety in the ham shack

Safety in the ham shack should be second nature to amateurs, but often it's not—with catastrophic results

Jack Darr and Alan James

Most of us don't think of it very often—in fact, not often enough—but safety precautions in the ham shack are doggone important. One careless move, and your call letters may be up for grabs (to someone who has more respect for the right way to do things). If people visit your shack, an unexpected slip, stumble, or fall might be more than just annoying or painful; it might slide someone into the hospital or undertaker's, and you into a law-suit.

Actually, there's nothing hard about safety precautions. Most are just good old common sense. Wire has insulation, power supplies have covers, transmitter cabinets have interlocks, chassis have ground posts—all for a purpose: to keep the careless or untrained out of danger. Many of the simple safety devices on commercial equipment can be added to homebrew gear with very little trouble. A few cheap signs and a little thoughtful housekeeping around your shack will probably take care of most hazards.

the protection racket

There are three classes of people you've got to think of: **You,** who spends more time in there than anyone; **your family,** especially the ubiquitous kid brother (or kid sister) who may sneak in while you're not around, and **casual visitors,** like girls (you remember, those high-pitched voices you sometimes hear on the air), or neigh-

bors, or just family friends; all may be the inquisitive type who just can't resist touching "all those cute little things" on the bench.

Certain standard precautions will keep the last two groups out of most trouble. Just keep all of the hot stuff safely boxed up in tight, well grounded metal cases. The only thing that could bite 'em then would be an open transmission line. Big, red-lettered signs labeling open voltages are a warning to anyone who can read, but—judging from the number of people who have gotten needless rf burns in one shack we know of—not everyone does. Complete insulated shielding is the surest protector.

Keeping all three classes of ham-shack inhabitants safe is your responsibility. Make a "safety inspection" of your own shack. Look for hazards you hadn't thought of, and clean 'em up. If you don't find at least a couple of things that need fixing, it'll be unusual.

no horseplay

Talking about visitors brings up a serious point: don't horse around with electricity! We all know the type of guy who hands other people charged electrolytic capacitors, or touches them with hot wires and that kind of stuff, just to see them jump. Guys who do this are idiots. It's not only painful, but can have sobering consequences. If you give someone a playful jolt and he turns out to have a weak heart, you may find yourself trying to beat an involuntary manslaughter rap!



"And there's the class II Idiot"

And, there's the Class II Idiot. This is the guy who comes up behind you when you're working on a piece of gear with high voltages exposed and yells "Boo!" or jabs you in the ribs. Only once did either of us ever tell anyone to leave the shack and never come back, and that was one of these numbskulls. Fortunately, the jolt he caused was merely painful, but it wasn't funny. Working on anything with exposed high voltages, you're naturally tensed up and ready to jump anyhow. If you aren't, you should be.

It's no place for you to show off, either. You'd best keep a healthy respect for the voltages and rf fields that abound. Don't try to prove you're a big smart operator who knows just how to play around live transmitters without getting killed. You only get one chance to be wrong.

During World War II, we knew a young smart-alec ham in the Signal Corps who knew "all about transmitters." He worked around high-power transmitters and wore a finger ring. We told him to take it off, but he wouldn't. One day he reached into a 1-kW transmitter to "make an adjustment" with the soup still boiling. You know what happened when that ring got into the high-intensity rf field: the tank coil made an efficient rf heater. He came out a heck of a lot faster than he went in. He suddenly wanted that ring off. After the big blister had healed up, he did get it off and never put it back on.

the shock treatment

The danger of electric shock is the ham shack's worst hazard. Exactly what can cause shock is not as well understood as it should be. You gotta know what to watch out for if you plan to keep your

beam from being lowered to half-mast.

This doesn't mean only novices, who might not know the different hazards. It also applies to the OM who's been lucky. He may have worked around the stuff so long without a jolt that he has gotten contemptuous. **Don't!**

How much "electricity" does it take to put you off the air permanently? Not your rig, but **you!** 2,000 volts? 5,000 volts? 110 volts?

Pick up a #47 pilot lamp. It draws 150 mA at full brilliance. Only **one-tenth** of that through your heart can kill you. Just 15 lousy little milliamps! You couldn't even let the bulb light up to a dull red glow before you would become part of a "ground system" yourself (large conductive object, buried deep).

Scare you? It should. It isn't voltage necessarily that kills; it's the current the voltage can push through your body. Even the "lowly" 117 volts of ac house power can be astonishingly deadly, given an easy access to your body (through wet hands, wet feet, damp cement floor, grounded water pipe or chassis, and so on). Not all shock deaths in ham shacks come from transmitter high voltage.

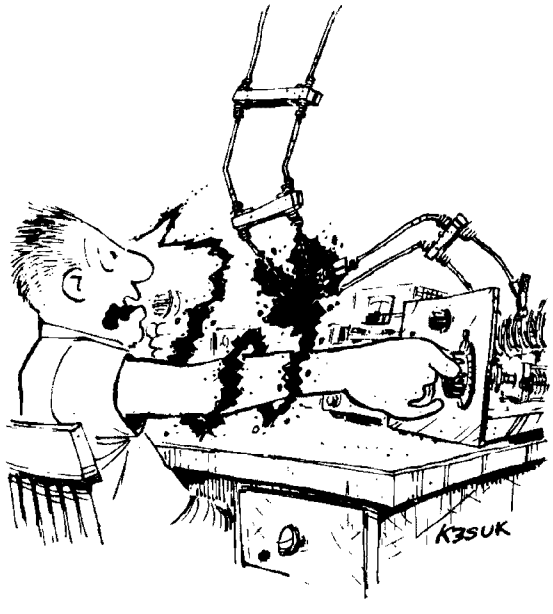
There's one thing a lot of hams forget. The ac line coming into your shack has two wires. The *black* one is hot, and the *white* one is grounded right outside the house, at the first utility pole, and at all the other poles—well grounded. The black wire, therefore, is always hot, with reference to **any** grounded object, and that includes **you.** You can light a 1,000-watt lamp from **one wire** (the hot one) merely by furnishing a ground for the other contact of the lamp.

In 220-volt systems, the black and white wires are both hot, and a **green** wire is the ground (called neutral or common). In three-wire "safety" cords on a lot of newer tools, appliances, and the like, the black wire is hot, the white wire ground, and the third (green) wire goes to the round pin of the AC plug—for a separate ground, usually to the metal case of the unit. However, don't put blind trust in color-coding! Even electricians make mistakes sometimes, and you might find wire-colors reversed.

So, if you happen to be grounded, and touch a hot side of the ac line with one hand (even with the other one in your pocket, as the old saw goes), **bang!** Standing on a dry wooden floor, you might get away with it. On a cement floor, on bare ground, or with any part of you touching a grounded object, look out.

The basement shack is dangerous because it usually has a cement floor. Even vinyl tiles or linoleum seldom help. Moisture works up between the tiles.

There are two things that will help. One: put down a thick rubber runner or mat on the floor near your gear. Use one-piece stuff, without cracks. Don't trust wood; it absorbs moisture. Two: wear shoes with rubber or plastic insulating soles. Caution: if the soles are of the sponge or foam type, and you've been walking through water, look out—you've lost your protection!



High-energy rf voltages have the disconcerting ability to reach out and burn.

A wooden-legged stool or chair is best for working or sitting. If you have a metal stool, put plastic cups on the ends of the legs.

If you're working on ac-powered gear, the first thing to do is pull the line plug,

and lay it on the bench where you can see it. Also, be sure you pulled the **right** plug. Many's the ham who has gotten the whop knocked out of him by pulling the wrong plug and then grabbing a wire in the wrong place!

hard-hitting dc

Hams seem to pay attention to the dangers of high-voltage dc in their rigs. Rightly so. It can be a dangerous source of electrical energy. Voltages range upward from 300 volts in receivers, up to 1,500-2,000 volts in transmitters. In well designed power supplies, shunting resistors across each electrolytic capacitor bleed off the voltage stored in the capacitors when power is turned off. They bleed it off **if** the resistors are good. You have no way of knowing, unless you check with a voltmeter. Don't put much faith in bleeders. A "shorting switch" may be provided to short high voltage to ground when the cabinet is opened.

If these devices work, okay. But the handiest gadget you can have for working around lethal electrical energy is a suspicious mind! Before you poke your little pink, uninsulated finger in there, check for residual voltages!

long-reaching rf bite

High-energy rf voltages have the disconcerting ability to reach out and burn you without your even contacting the circuits carrying them. Consequently, the inside of live transmitters, and the feed lines from them to the antenna, are places to keep your distance from. Coaxial cable is safest for the feed line, but it isn't feasible at some frequencies and with some kinds of transmitters.

If you have a high-power rig with an open-wire antenna lead, you can put this line too high for anyone to reach. Or you can put a plastic cover over any exposed antenna wires, switches, etc. The antenna line itself can be ordinary automotive ignition wire. The insulation on this handles up to 25,000 volts, which will safely contain even a California kilowatt. It's even available with "pretty" bright-red plastic insulation, to dress up the shack.

Under no circumstances use a metal screwdriver to make adjustments in high-power rf circuits. This is a real shock-hazard. Besides, metal screwdrivers detune rf circuits. Use insulated tuning tools or shaft extensions.

protection from ufo's

That pretty well takes care of you, if you use your head. Now, how about the rest of the world, especially Kid Brothers and such? KB's are good at sneaking in and turning the whole thing on to show visiting



Kid brothers are good at sneaking in and turning the whole thing on to show visiting buddies.

buddies. In fact, even a non-ham Pop isn't immune to that kind of temptation: "Look at this thing my kid built! How about that? Lemme show you how it works!"

There's one way to make sure the Unlicensed Foolish Operator (UFO) can't get into trouble. Wire the rig to a separate ac power supply, and put a master switch on it. When this is turned off no power reaches the rig at all.

This is easy to do, with heavy Romex wire and "surface" fittings. Use an enclosed switchbox with fuses inside and a lever-operated double-pole switch to open both sides of the line. These switches have

provisions for a small padlock to be slipped over the lever. The thing will be safe unless Kid Brother finds the key. So, keep it with you. If there is a spare, give it to Mom for safe-keeping; she won't let KB or Pop have it!

safety and showmanship

Back when most ham gear was homebrew (there weren't any kit companies), there was a lot of exposed wiring. Haywire "lashups" were common—wires going everywhere. Nowadays, a little ingenuity can make the shack look as smooth as a broadcast station. This is not only much better looking, but a whole lot safer for everyone concerned. It keeps the hot circuits and wiring covered up.

How? It's easy. Just put all the gadgets into boxes or behind panels. You can buy metal boxes in any size you can imagine. Power supplies, antenna couplers, preamps, modulators, or anything else, can be built into these boxes. If you have some unit that can't stand the added capacitance of a metal box—say, an antenna coupler—use a non-conductive box: masonite, bakelite, acrylic, polystyrene.

With nice professional-looking, neatly lettered panels on the front, your boxes can dress up the whole place! You can find knobs in any color or size that looks nice. Lettering has always been the hardest part. Now, you can buy lettering decals in black or white, in almost any size. They're simple to put on, and give the whole thing a neat touch. Also, you'll be surprised how much easier (and safer) the equipment is to use when all controls are labeled!

If you have a chassis so goodlooking you hate to cover it up, use a clear plastic panel. Quarter-inch stock can be bought in almost any size sheet you need. (Try aircraft supply houses; plane windows are plastic. Or look around the boneyard at the airport; wrecked planes always have "lots of good parts.") Use black lettering on the clear plastic panel. For a final touch, turn the pilot light so it illuminates the chassis when it's turned on!

With the stuff enclosed in boxes and panels, the shack looks a lot better. You also keep it safer, and the worst trouble a visitor is likely to experience is a bawling-out from you for fiddling with all those pretty knobs!

safe and sane construction

Commercial and kit gear is pretty well safety-ized; homebrew gear is something else, but it shouldn't be. When you build a new transmitter or receiver, if there are any voltages present higher than about 25 volts dc, take precautions. There are a few simple steps that make any rig safer to use and work around.

For one, be sure the ac power coming to the switch and power transformer of the rig is well protected. Mount a strong three-terminal strip just inside the chassis, and solder the ac cord solidly to the lugs, along with the leads for the power transformer primary and the switch.

In transmitters, there are several controls to adjust for proper tuneup or loading. The shafts of some of these must be "hot"—for example, the small trimmer capacitor across the final tank in the antenna circuit. When you mount such controls, situate them so you can use insulated extension shafts. Then you can make tune-up adjustments from outside, in perfect safety. You can get plastic tubing 16–18 inches long that'll fit over a 1/4-inch shaft. All you have to do is slip the open end over the shaft and fix it there with a dab of epoxy-resin cement. Leave about an inch of the tubing sticking out, if the control works easily, and you won't even need a knob.

A new piece of gear usually starts out in breadboard style, even in fancy electronics labs. Temporary wiring, if done sloppily, can be dangerous. To keep it safe, use well insulated wires, and make sure all power connections are tight (so they can't work loose and touch things they shouldn't).

There's one good way to connect two wires together temporarily. Take a 6-inch piece of spaghetti large enough to slip over the insulation. Slide the spaghetti over one wire. Strip about 1/4-inch of insulation from the end of each wire, and solder a

smooth lap joint. When the solder sets, slip the spaghetti over it so the joint is right in the middle. When you're ready to put the thing into permanent form, slide the spaghetti back, and a tap with the hot soldering iron will open the joint. You can even put the spaghetti "back in stock" unharmed. No waste at all!

The lap joint is also useful for temporary connections in the apparatus itself. Tack the wire end to the terminal lug solidly. Don't make a hook in the wire; let the solder do the holding. The joint will be safe and solid, but easily disconnected when you get the final design done.

Exposed multiple-screw terminal strips on the back of a piece of gear might not seem particularly dangerous since they'll be well out of the way—or, will they? Try dropping a pencil or something down behind and then reaching for it. (Voice-Of-Experience stuff again!) All those terminal strips can be covered up, either with plastic covers bolted in place or with a wide piece of cloth-backed adhesive tape. The tape is a good temporary precaution during construction, as well.

When you make interconnections between units, make them safe. The best way is to make them plug in. A good source of cheap plugs and sockets is a junked TV chassis; octal tube sockets and old tube bases can give you enough plug/socket combinations to last a long time. One hint: when you hook these up, be sure the voltage **source** is connected to the **female** plug! This way, the contacts are out of reach. Never have voltage on a male plug, with its exposed pins; this is courting trouble.

Safety in the ham shack is really simple and easy if you go at it in the right way. All it takes is a little common sense, some insulation and decent housekeeping. Keep the hot stuff well covered up, so the ignorant or uninitiated can't get their fingers into it. Keep the floors cleared so no one trips and breaks something (of yours or theirs). Practice looking for possible hazards, and then take precautions to make them harmless.

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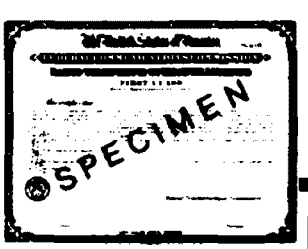
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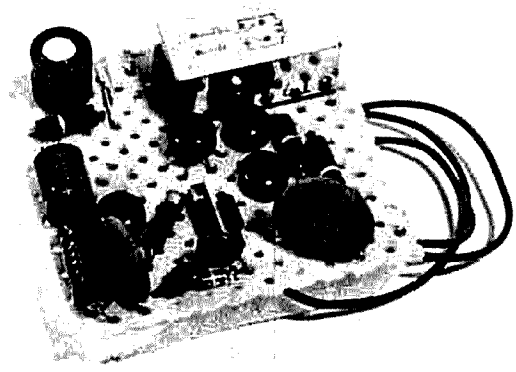
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versatile ic vox

The many faceted μL914 IC in a simple but effective microphone control circuit

John J. Schultz, W2EEY, 40 Rossie Street, Mystic, Connecticut 06355

Using an inexpensive integrated circuit for stage functions, this vox unit can be added to any transmitter. Operation is particularly stable and adjustment simple. If an antitrip stage is not needed, the circuit requires only two IC's and one transistor.

An effective vox circuit should have

1. Good sensitivity.
2. Positive relay switching.
3. Good antitrip action.
4. Independence of the controls for over-all sensitivity, antitrip sensitivity, and time delay.

The unit described may not be the last word in vox performance, but considering its relatively simple circuit, it performs much better than many elaborate tube or transistor units. The switching action is positive, and it has none of the transient problems associated with many other circuits, particularly those using tubes. Its construction and power supply requirements are straightforward. It can be easily incorporated into an existing transmitter or built as an outboard unit for mobile or portable service.

The heart of the unit is several Fairchild μL914 integrated circuits. This versatile IC costs less than a dollar from major supply houses and is fast becoming one of the most popular units for amateur use. In fact, you might say that it is the 6V6 of integrated circuits. Part of this popularity derives from its basically simple internal circuit, **fig. 1.** The circuit is certainly simple compared to many other IC's, but this very simplicity allows it to be used, by varying its external connections, for a wide variety of uses ranging from switching to amplifying circuits. Packaging four transistors inside the unit allows compact and economical circuit construction.

The vox circuit illustrates the wide variety of uses for the μL914. Even if you don't build the vox, the circuits may inspire ideas for other applications.

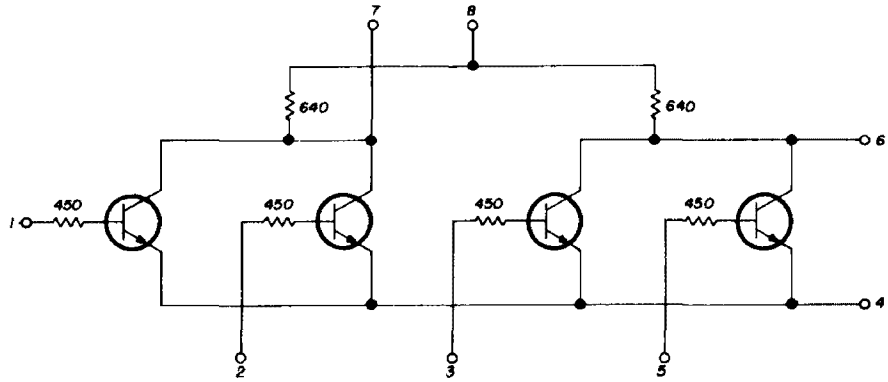


fig. 1. Internal circuit of the μL914 integrated circuit.

the circuit

The complete vox circuit is shown in **fig. 2.** Three basic stages are used, each containing one μL914—an input amplifier, inhibit gate, and delay generator.

The audio input circuit is conventional. If you compare the terminal connections of the μL914 in **fig. 2** with the actual μL914 internal circuit of **fig. 1,** you will see the amplifier consists of two tandem grounded-emitter stages. The capacitors at pins 1, 5 and 6 are not adequate for good speech reproduction (they should be 20 to 25 μF), but this is not important for this particular application.

The audio input can be taken from the first audio (microphone) amplifier stage in the transmitter, if sufficient level cannot be obtained directly at the microphone connection. Alternatively, if this is inconvenient, another μL914 amplifier stage can be added to the vox, with the sensitivity potentiometer connected between the two stages.

The second μL914 stage is an inhibit gate or switch for antitrip action. It prevents microphone pickup from the station loudspeaker from activating the vox circuit. The antitrip reduces the vox's sensitivity to speaker signals by reducing the over-all vox gain. The second μL914 performs this function in a unique manner.

If the terminal connections for the μL914 are followed, it will be seen that one transistor is an ordinary grounded-emitter amplifier. Another transistor is a shunting element, with its collector and emitter across those of the first transistor. A positive control voltage applied to the base of the second transistor decreases its collector-emitter resistance and shunts the output.

The control voltage is developed by simply rectifying part of the speaker voltage. As shown, a transformer (miniature transistor type) matches the speaker impedance. If an audio connection can be made at a high-impedance point, the transformer will not be necessary, and connection can be made directly to the antitrip level potentiometer through a 1-μF capacitor.

The entire vox unit can be considerably simplified if the antitrip feature is not required. In home stations where the microphone gain is run fairly high and loudspeaker operation is used, antitrip is required. In situations where the microphone gain is run low

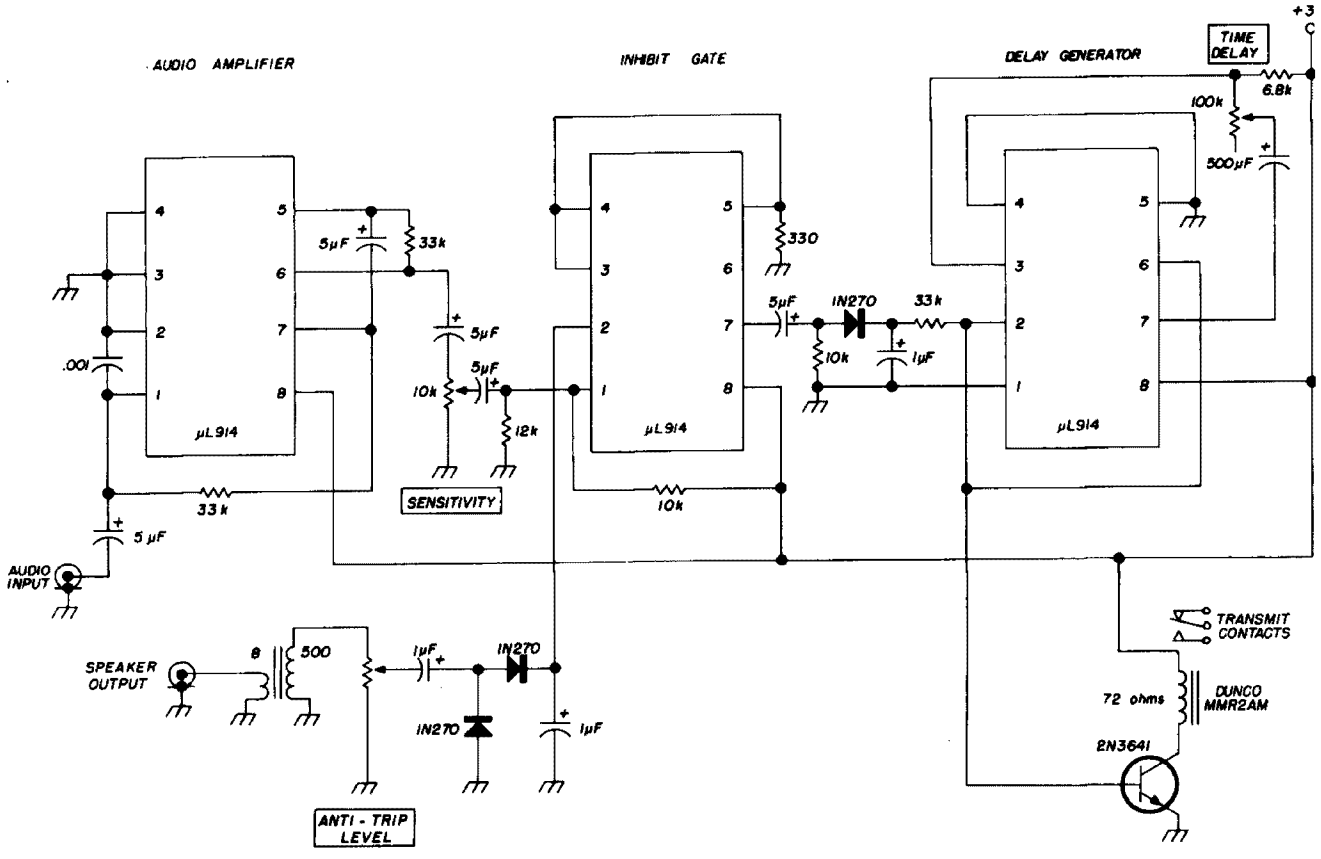


fig. 2. Schematic of the IC vox; each integrated circuit is a µL914. The relay is a Dunco MMR2AM.

and the microphone held closely to the lips, antitrip is often unnecessary, even with loudspeaker operation. If antitrip is not needed, the second µL914 can be eliminated, and the arm of the 10k sensitivity potentiometer connected to the left side of the 5-µF capacitor that is connected to terminal 7 of the second µL914.

The third µL914 is connected as a form of monostable vibrator. A positive input pulse produces a positive output pulse. The rectifier circuit between pin 7 of the second µL914 and pin 2 of the third µL914 provides the positive input pulse. The output pulse remains positive for a period after the input pulse has decayed; this is determined by the time constant provided by the 100k potentiometer and 500-µF capacitor. A maximum delay of about 3 seconds is possible, although this can be increased as desired by increasing the capacitor value.

The output pulse from the delay generator controls the 2N3641 switch, which in turn activates the relay coil. The relay contacts are connected across the transmitter push-to-talk circuit contacts. The relay can be almost any low-coil-resistance type that works on 3 to 6 volts.

construction

Actual construction depends upon how the vox unit will be used. If used as an outboard unit, it should be placed in a shielded enclosure. In general there is nothing critical about the wiring.

The photo shows how the components can be easily assembled on a piece of Vector board. The potentiometers can be either mounted on a separate control panel, or miniature PC types can be mounted directly on the Vector board if frequent readjustments are not made. In fact, for a stable situation with specific equipment, they can be replaced with fixed resistors. A suggested parts layout is shown in the photo, but any similar arrangement can be used. The µL914 terminal

numbering in **fig. 1** should be carefully noted. The index mark denotes terminal 8, and not terminal 1.

Other types of integrated circuit "dual 2-input NAND/NOR gates" may be usable in place of the μL914, but the specifications for their internal values should be carefully checked to see that they generally correspond to the μL914.

The vox may be battery powered from D cells or operated through a voltage divider circuit from a low-level dc source in the transmitter. If battery operated in a mobile station, a zener voltage regulator is desirable to prevent voltage fluctuations from changing the time delay.

summary

Adjustment is quite simple. If the antitrip stage is not used, the sensitivity control is set for the microphone level at which you want the transmitter to be turned on. The time delay is adjusted for the desired pause period between words until the vox disables the transmitter. If the antitrip feature is used, the procedure described above should be followed with the speaker volume turned down. The speaker volume is then turned up and the antitrip level adjusted so the speaker output does not activate the vox. Control interaction is nil.

A further interesting use to which the vox unit can be put is to add cw operation to a ssb-only transceiver. A keyed audio tone is fed to the transmitter audio input and the vox unit. The vox then acts as a cw break-in circuit with an adjustable time delay. One important precaution that must be observed, however, is that the tone be of a single frequency and as pure as possible. The output of any tone oscillator used should be checked on an oscilloscope. Through suitable amplifier circuits, the keyed tone can also serve as a sidetone for monitoring the keying.

A final word of caution concerning handling the μL914 is in order. Although it will withstand considerable environmental extremes, the μL914 can be easily ruined by excessive terminal heat or incorrect supply voltages. The same precautions applicable to all transistor circuits should be observed.

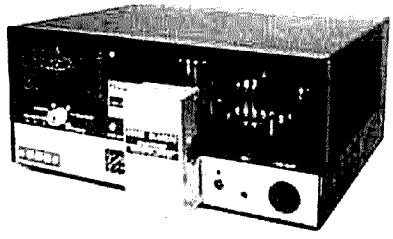
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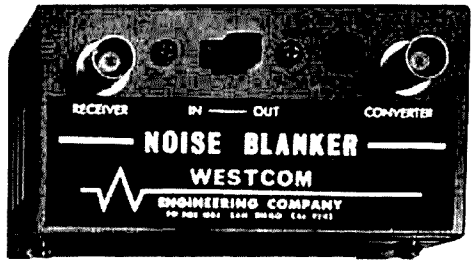
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propagation predictions for march

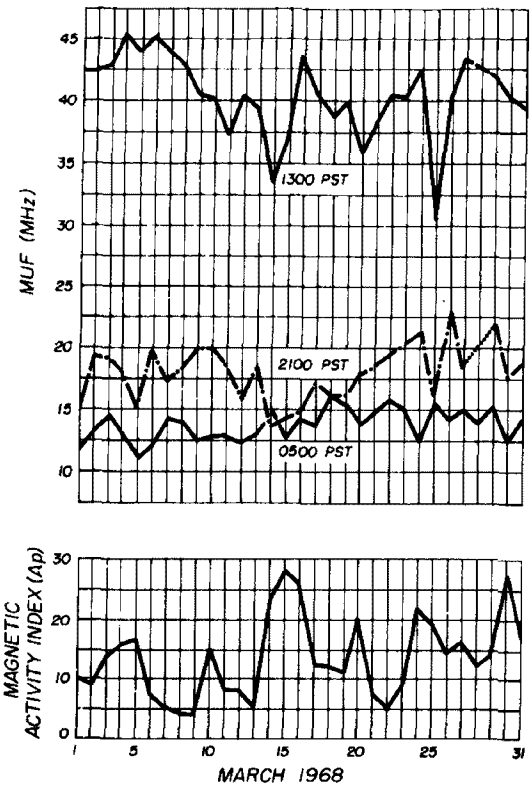
Victor R. Frank, WB6KAP, 12450 Skyline Boulevard, Woodside, California, 94062

Let the reader beware; I am in the process of modifying the computer program used to derive the maximum range charts for this column. The first step (incorporated in the February column) was the conversion of distances from kilometers to statute miles (previously done after the chart was drawn). The only change you will notice is that the minimum range is now 1000 miles instead of 1000 km (625 miles). The problem: new "refined" formulas may cause a greater change in predictions between months than is actually present. Next month 160 meters will leave the charts.

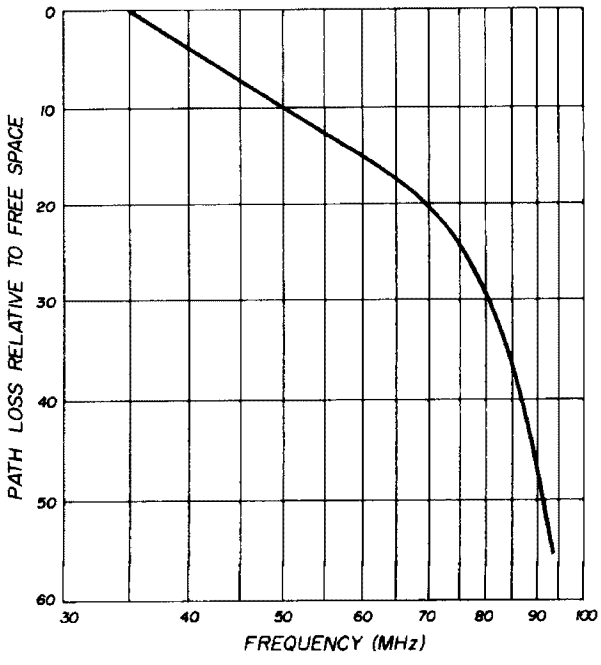
Predictions of sporadic-E activity may deservedly cause you to entertain some reserved skepticism. Looking back on December 1968, not only were there no two-meter sporadic-E openings reported (in the Northern Hemisphere), but six meters didn't fare well by sporadic-E either. However, ZK1AA in the Cook Islands moved his beacon to 50.100 MHz, and I have reports of it being heard in Southern California on December 5, 7, 9, 10, 21, 22 and 29, generally between 1400 and 1530 or between 1830 and 1930 pst.

maximum usable frequencies

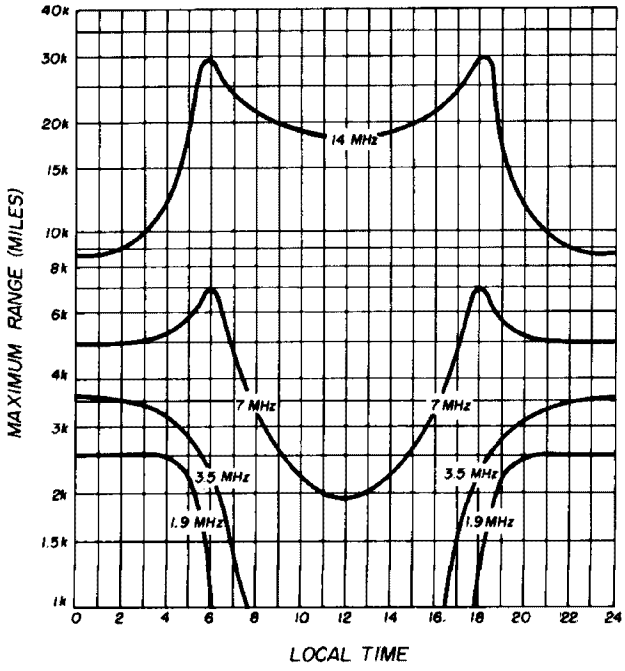
It is expected that solar activity will continue its very slow descent from its peak in 1968. The small change of smoothed sunspot numbers from last March (103 for March 1968, 98 for March 1969) may only be noticed by 50 MHz operators, and an increase in the number of ionospheric disturbances **could** result in even more openings. The most likely 50-MHz openings during March and April



1. Maximum usable frequencies (F2 layer) scaled from vertical incidence ionograms taken at Point Arguello, California between March 1 and March 31, 1968 compared with magnetic activity.



2. Variation of path loss with frequency for transequatorial propagation between Oahu, Hawaii and Raratonga, Cook Islands, during March 1968.



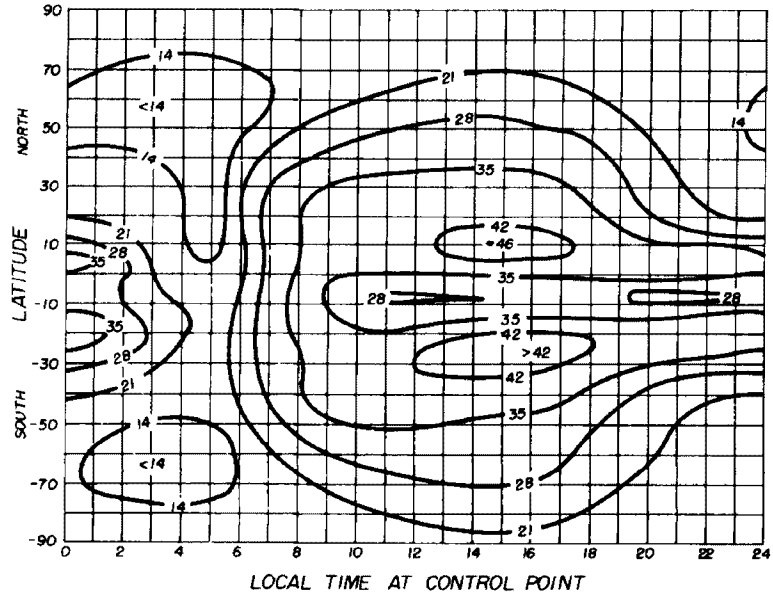
4. Maximum range due to absorption and noise vs local time from 38° N. Latitude to the north.

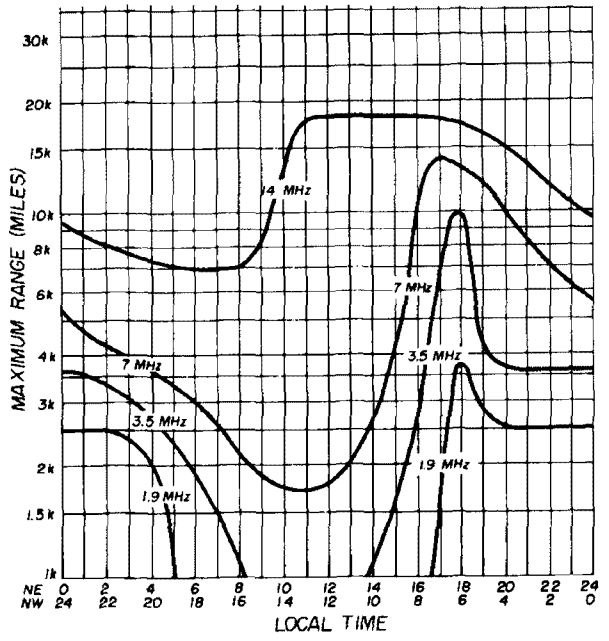
will be by transequatorial forward scatter (for Southern United States) and by aurora (for Northern United States) during the evening hours.

Maximum usable frequencies during March 1969 will probably be within a few percent of those of March 1968. I have scaled vertical-incidence ionograms taken at Point Arguello, California (35.5° N. latitude) during March 1968 to determine the muf for the maximum one-hop distance for a path with the control point at that latitude. The muf's at 0500, 1300 and 2100 local time (pst) are indicated for each day of the month in **fig. 1.** The median muf's predicted by ITS for a 4000 km path were 14.8, 36 and 20.5 MHz, respectively.

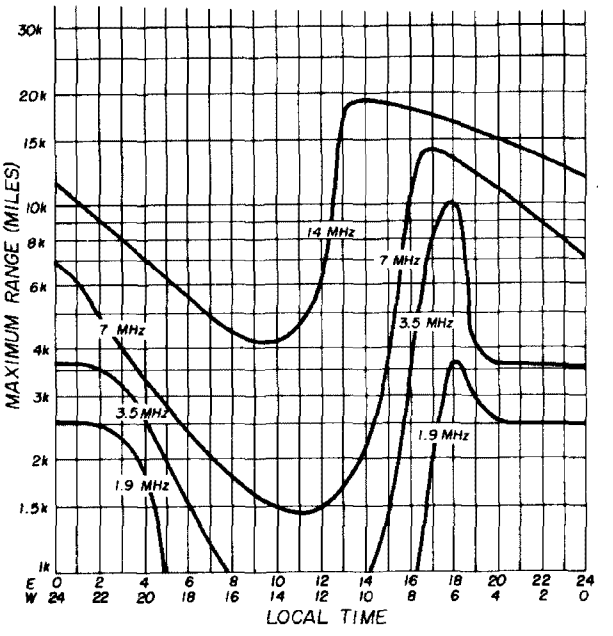
It appears that ITS overestimated nighttime muf's and underestimated daytime muf's. The predawn minimum of F2-layer critical frequencies occurs near 0500. The

3. Time chart of predicted median muf for March 1969 based on 90° W. Longitude.

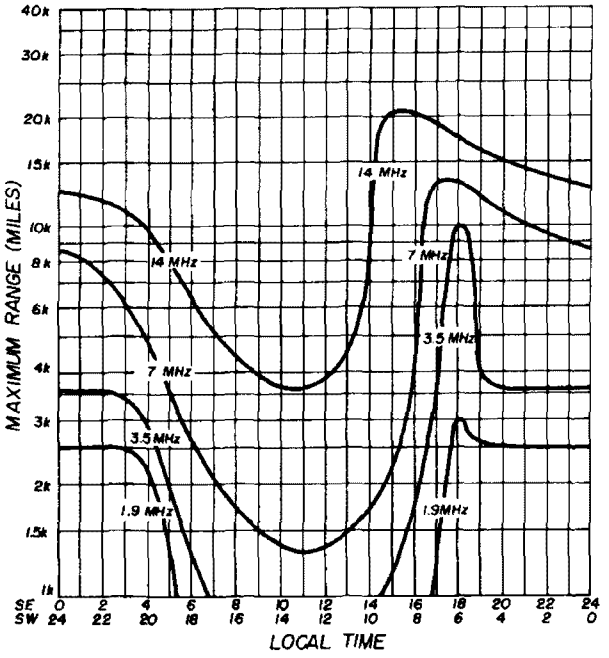




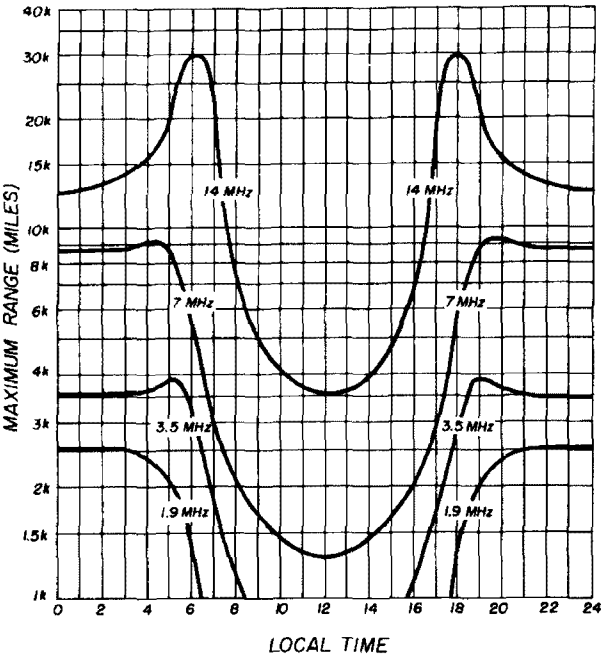
5. Maximum range due to absorption and noise vs local time from 38° N. Latitude to the northeast (top time scale) and the northwest (bottom time scale).



6. Maximum range due to absorption and noise vs local time from 38° N. Latitude to the east (top time scale) and to the west (bottom time scale).



7. Maximum range due to absorption and noise vs local time from 38° N. Latitude to the southeast (top time scale) and the southwest (bottom time scale).



8. Maximum range due to absorption and noise vs local time from 38° N. Latitude to south.

how to use these propagation charts

1. To find the maximum usable frequency for F2-layer propagation for distances of 2500 miles or more in any direction, read the frequency at your control point from the muf time chart. Your control point is 1200 miles away from your station in the direction of the propagation; this is about an 18-degree difference in latitude for a north-south path, or 1½ hours difference in time for an east-west path. The muf time chart may be treated as an muf contour map in the longitude interval between 45° W. and 135° W. As such, each hour is the equivalent of 15° of longitude. A map drawn to the same scale could be overlayed and positioned to the right or left to show the variation of the muf contour map with time.

2. Over any particular path involving more than one hop, the path is the lower of yours *and* the other *station's control-point* muf. Curved lines may be drawn on the overlay representing the great circle path, as found from a globe or "Ionospheric Radio Propagation," printed by the U. S. Government Printing Office.

3. To find the maximum propagation distance as limited by ionospheric absorption and atmospheric noise, refer to the maximum range charts for the directions you wish to work. Note that tha time scales are reversed for westward propagation. These curves are based on unity signal-to-noise ratio in a 6-kHz bandwidth with *100 watts output* power and antenna gains (over an isotrope) of 6 dB for 20 meters, 0 dB for 40 meters, and —6 dB for 80 meters at each station.

The muf time chart was derived from "Ionospheric Predictions" for a longitude of 90° W. These predictions are published monthly by the Institute for Telecommunications Sciences (ITS), Boulder, Colorado and available through the U. S. Government Printing Office. The maximum distance curves were derived from consideration of atmospheric noise levels (from CCIR report 322) and calculated path losses at fixed distances in each direction from 38° N. latitude. Some minor differences in maximum range would be noted due to change in absorption for stations located between 26° N. and 50° N. latitude. Somewhat greater ranges would be expected over paths further from tha subsolar point (more northerly latitudes).

The predictions given in this column are for median conditions. On any particular day, muf's may be as much as 10% higher or lower than the median. Absorption and noise levels, particularly on the lower amateur bands may be as much as 10 dB different from the median. Residential noise levels (from electrical lines, appliances and vehicular traffic) may, and frequently will, be tens of decibels stronger than atmospheric noise.

highest muf's occur near 1300 or somewhat thereafter. Prime operating time (for those who work days) is near 2100, which also happens to be in the middle of the transequatorial scatter period (not evident on soundings this far north). Note a decrease in daytime muf's as the month progresses, and an increase in nighttime muf's. The lower curve is of the magnetic activity index Ap.

effect of magnetic storms

Magnetic storms occured between March 3, 0700 and March 6, 0100; March 9, 2300 and March 11, 1200; March 14, 0200 and March 17, 2000; March 23, 0700 and March 27, 1100; March 29, 0200 and March 31, 0000; and March 31, 0300 to April 1, 2300 (gmt). The most obvious effect of the magnetic disturbances was to lower the 1300 and 2100 muf's. The predawn muf's were affected to a much smaller degree. One effect noted on the nighttime ionograms during the storms at the middle and end of the month was spread-F, a time-delay spread echo from ionospheric irregularities, i.e., the ionosphere no longer behaved like a smooth reflecting surface.

transequatorial propagation

Transequatorial scatter propagation involves oblique reflection from F-region irregularities near the magnetic equator. Transequatorial propagation (TE) may occur at frequencies at least twice as high as the muf for regular ionospheric refraction. The signal is often spread in time delay and undergoes Doppler shifts that correspond to reflection

from equatorial F-region irregularities drifting east-west as fast as 330 mph. **Fig. 2** shows the variation of minimum path loss with frequency for TE propagation between Oahu, Hawaii, and Raratonga, Cook Islands, during March 1968. Peak signal levels occured between 2100 and 2200 local time. For reference, well-equipped amateur radio stations could probably communicate with a path loss of 50 to 60 dB in excess of free-space loss at this distance.

In view of the preceding, and infrequent propagation at 108 MHz, it appears that 144-MHz TE is somewhat beyond the range of amateur radio capabilities. However, the same has been said of 220-MHz meteor scatter and 432-MHz aurora.

seasonal changes

As the vernal equinox (March 21) approaches, the effect of geomagnetic control on the F2 layer becomes more evident in the muf time chart, **fig. 3.** At 90° W. longitude, the muf contours are centered on the geomagnetic equator, about 9° to the south of the geographic equator. Seasonal changes from earlier months are:

1. Higher noise levels; 80- and 160-meter noise levels at midnight are expected to be up an average of 8 dB during Spring.

2. Lower daytime F2-layer muf's in north temperate latitudes, especially during morning hours, possibly higher F2-layer muf's during midafternoon in tropical latitudes.

3. Higher nighttime F2-layer muf's with fifteen meters being open as late as 2200 local time and twenty meters staying open all night during the latter part of the month for southerly paths.

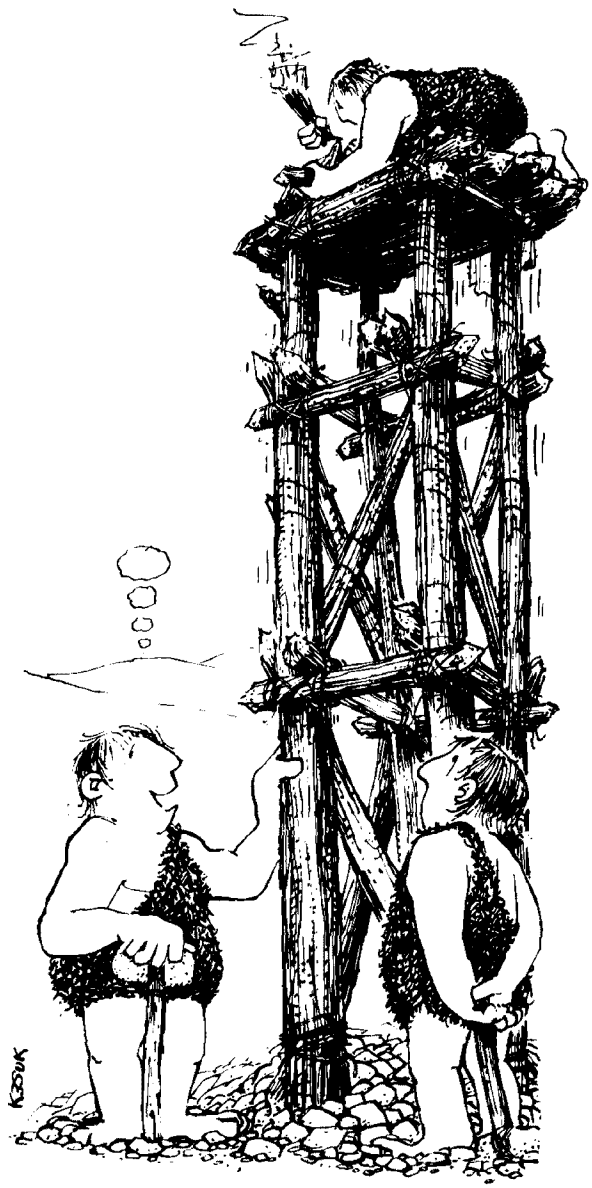
4. Higher absorption during daytime in the Northern Hemisphere.

5. Polar paths will be increasingly difficult on 21 and 28 MHz as the muf contours move southward.

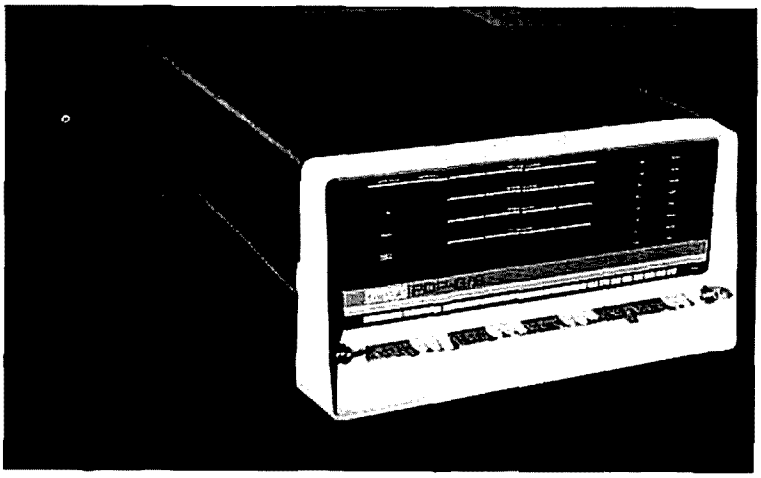
Optimum conditions are expected for contacts between Northern and Southern Hemispheres.

ham radio

a look at those days before spark... b.s.



. . . so I sez to him look . . . you want your signal to get out, you gotta get it up in the air . . .



Desk top computer that *could be easily adapted* to amateur use*.

computers and ham radio

This is probably as far-sighted as you can get, but it's a challenge to the imagination

Louis E. Frenzel, W5TOM, 4822 Woodmont, Houston, Texas 77045

Have you noticed the publication explosion that's taken place recently in electronics? Next time you're at your favorite parts supply house take a look at the book rack. The number of books and periodicals devoted to solid-state technology is staggering. You really have to dig way back into the dust bin to find an article on the application of a 6AK5 in a vhf preamp, for example. *Indeed, it seems rather pointless* to fool with **any** vacuum tube unless it's required for developing large amounts of rf power. We may as well face the facts: the vacuum tube is going the way of such *quaint items as the razor strop and* the silver fifty-cent piece.

It's hard to find any electronic circuit that hasn't been adapted to solid-state devices. For example, a recent issue of one popular amateur electronics magazine had a total of nine articles involving the use of solid-state devices but only one that used vacuum tubes. So it's a healthy sign: amateurs certainly aren't dragging their feet.

One area of technological advancement in electronics that hasn't made much of an appearance in the amateur field is the computer. During the past ten years the computer has developed into a powerful tool for business, industry, education, and the government. All our lives are affected by the computer. If anyone doubts this, all he has to do is try to get a gasoline credit card cancelled. One fellow spent about five bucks in stamps and telegrams trying to do this, and he finally gave up. Business

The Digital Equipment Corporation's PDP-8/S computer sells for under $10,000. The basic system includes 4096 words of 12-bit memory and is capable of real-time, on-line computation.

is so automated these days it's somewhat frightening. However, I have no doubt that the future will bring forth a computer that will not only read handwriting, but will sass you right back in your own language. In fact, right now it's possible to talk to computers in a rather crude way—but more of this later.

The big question is, what effect, if any, will the computer have on amateur radio? I'd like to theorize a little and consider some of the implications of the computer. Computers are already being used in amateur radio. For example, some ham organizations and equipment manufacturers use them to take the drudgery out of bookkeeping, and there are many other applications. This is pretty indirect, but valid.

There's been a lot of mystery surrounding the computer, mostly in the minds of the lay public. You hear all sorts of remarks such as "thinking machine," or "electronic brain," etc. Well, first of all I'd like to set the record straight. The digital computer is **not** a thinking machine, nor is it any kind of brain. As a matter of fact, the computer is a charter member of the idiot squad. It's stupid. It can only do what it's told. Furthermore, if it's not told correctly, all kinds of problems can develop. For example, the IBM 7090 general-purpose computer costs something like $500 an hour to operate, depending on the amount of peripheral equipment. If the machine isn't programmed properly it's possible to run up a pretty fair-sized tab. However, these machines have time-limit instructions in their soft-ware to circumvent loops or system-oriented problems.

Stated simply, a digital computer is nothing more than a sophisticated calculator with a memory, input/output (I/O) and peripheral equipment. The I/O consists of some means of communicating with the computer (input) and a means of obtaining computation results (output). Input could be punched cards, magnetic tape, or even a graph, depending on the machine. Output can consist of printed listings, cards, tape, or graphs.

You must learn a new language to communicate with the computer. Because these machines use on/off circuits, ordinary numbers as we use them have to be translated into the binary system. There are many books on binary arithmetic, so I won't go into an explanation.

However, it's pretty clear that, if the machine can be in either an "on" or "off" state, it can't very well handle numbers whose base is ten. Furthermore, the digital computer does nothing more than the four basic operations of arithmetic: add, multiply, divide, and subtract. And this is where the programmer comes into the picture. It's his job to translate complex mathematics into combinations of these basic operations.

computer development

The computer industry is young, and in the past ten years fantastic changes have taken place. Computers have undergone the same technological changes we have seen in ham gear. Early computers used relays, then tubes. Most recent machines are fully transistorized, and the newer units use integrated circuits. The use of integrated circuits has drastically reduced computer costs. You can now buy a complete computer for under $10,000. The $5,000 mark has just been made, and no doubt the future will eventually see a $1,000 machine. Already several small educational computers are selling for several thousand dollars. When we start talking in the lower price ranges, the feasibility of a computer in the ham shack doesn't sound too fantastic.

Just how could the computer be used? Since computers work exclusively with codes and pulses, their greatest application appears to be in the area of CW and RTTY. I'll list a few of the possibilities, keeping in mind that the computer is a versatile device with an extensive memory and can be programmed to do almost anything.

cw keyer

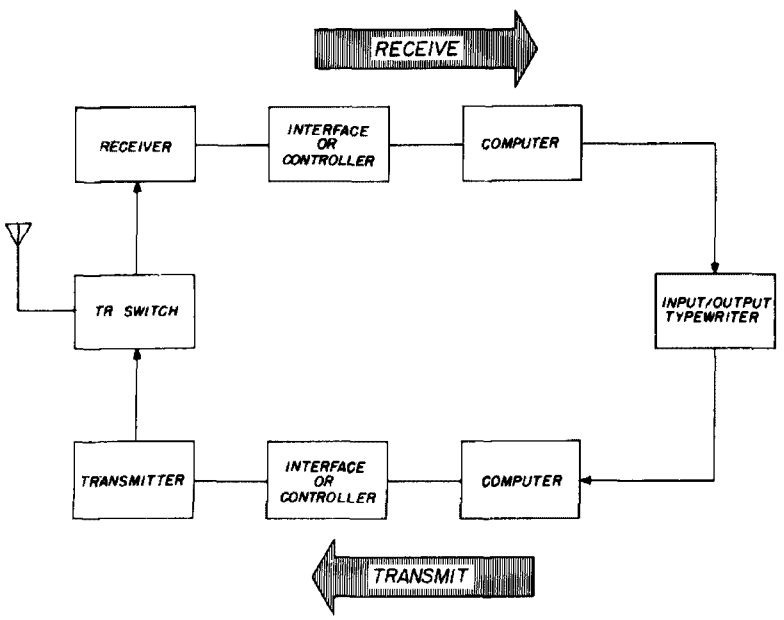
The computer could be used as an automatic CW keyer. The I/O typewriter would be used, as in teletype operation, to send the letters and numbers. The computer program would recognize the code generated

by the I/O typewriter and would, in turn, generate the appropriate dots and dashes to drive or key the transmitter. The program could be written to include a speed-selection feature wherein the desired output speed would be selected before sending by typing in the desired speed. The control program would recognize the speed and set up circuits to generate code at the selected speed. Naturally, this feature would be limited according to the operator's typing speed.

goodbye code practice

With the help of some special external circuits, the computer could also be used to receive conventional CW signals, interpret them and type them out on the I/O typewriter. The external circuits are needed to compensate for the wide variations in speed, spacing, and dot-dash length ratio. This is really a beautiful application since it would ensure a perfect, typed copy for every radio contact with no operator CW skill required.

fig. 1. Block diagram of a computer controlled amateur station.



The use of the computer merely as an automatic keyer doesn't really tax its capabilities at all. The computer's extensive memory would permit storage of complete words and sentences that could be called for by the operator from the keyboard. It would make the ultimate CQ wheel. In fact, entire conversations could be canned, including CQ and the initial response with name, signal report and location. All information would be stored as part of a master control program before the contact. Then the operator could call for this information when needed by simply typing in the proper message on the I/O typewriter. How about that? Almost takes the joy out of living.

rtty

Because of the wide versatility of the computer's logic circuits, it could also generate standard RTTY signals without additional hardware. Only a special program would be needed. Of course, the fsk or afsk modulator would be external. Having a computer in the ham shack would mean a total CW-RTTY capability in one package. And with the proper control program, the computer could perform CW-to-RTTY and RTTY-to-CW conversions so that the two modes could cross-communicate.

automated log-keeper

The computer is a willing worker and is quite adept at such routine jobs as book-

keeping or record keeping. With proper programming, it could perform the job of an automatic log keeper. During a computer contact, the operator would enter pertinent data such as dates, times, power, frequency and other required log information via the I/O typewriter. During the contact, this data would be stored along with that accumulated during the contact such as name, location, signal report, etc. All data would be stored during the day's operation. At the end of the day, the operator would simply call for a log printout, and the I/O typewriter would type out all log data in a predetermined format under the direction of the control program. The log printout could be filed, as required by law.

qsl monitor

If the computer's memory system was extensive enough, it could store information from the callbook, and upon completion of a contact it would automatically type out a QSL. Oh—too much.

contests

The automatic capability and record-storing feature of the computer would really be a boon to contest operators. In fact, the computer could conceivably carry out the whole contest alone if programmed properly. It would send and answer CQ's and exchange contest information. It could keep track of the station's call, contact number, and section worked and do cross-checking and log keeping. When it was all over, the operator would call for a log dump on the I/O typewriter. Out would come a neatly typed, complete log of all contest activity. Fantastic? But entirely feasible. Contest scores by computer should be sky-high.

Carrying this a bit further, when you're in your car on the way home from work the night of the contest, you could push a button on the dash and get the whole shebang into operation precisely at the moment of contest start. Then you could go into the house, take a leisurely shower, sort out your mail (who needs mail with computers around?), have dinner, read the latest issue of **ham radio** (which by this time would be a printout) and wander out to the shack to see what's going on.*

as a mechanized operator

The computer can also be used to control and monitor the operation of other equipment in the shack such as the receiver and transmitter. In fact, if proper circuits in these units were made available, the entire station could be set up from the computer. The transmission mode (CW, RTTY) and frequency of transmitter and receiver could be set up by just typing in the desired information.

Computer output pulses would control circuits in the receiver and transmitter to set the mode and frequency desired. Another example would be where the computer would recognize the location of your man during the contact, compute the distance to the station and its local time, then type out this information for you. The bearing angle would also be calculated under computer control.

the qsl bureau

The problem of exchanging QSL's could be simplified and improved if put under a computer system. One large central QSL bureau could be established with a computer to keep track of incoming and outgoing cards. Instead of the usual colorful printed QSL's, confirmations could be a punch card. While not as individualistic and colorful, the confirmation would be just as valid and easier for a computer to handle. Much of the usual paperwork headache associated with QSL bureaus could be delegated to the computer.

I've only mentioned a few items here for consideration. I think they illustrate the point well. But by no means do they represent the entire capability of a computer for ham applications.

Many of you will read this article with some skepticism, and I don't blame you. It does all sound a little farfetched. The amazing thing about it, though, is that it's

* You'd better have on a padded suit, because the rascal could very well reach out and backhand you for not kissing it goodbye before leaving for work that morning. **Editor**

all entirely possible today with existing technology. All it takes is money and initiative.

A small computer could be purchased outright for about $10,000 to $15,000. It could then be interfaced with the other units as shown in **fig. 2.** The nature of the interface electronics would depend on what you want to do with the computer. Then comes the tough part—writing the control program. Because of the versatility, you could program almost anything. For about $15,000 and a little time, study and effort, you could own a complete computer-controlled station, TODAY! Who will be the first?

One nice feature about a computer-controlled station is that when you're not hamming you can use the computer for other applications. Bear in mind the computer is a general-purpose unit that can be programmed to do almost anything. You could use it to maintain a budget, keep your bank account up-to-date, keep track of charge accounts and figure your income tax. You could use it to estimate your chances with a given horse race or a stock on the securities market, and your wife could use it for keeping track of addresses, telephone numbers and Bridge scores. The applications are unlimited. Perhaps now we could justify the cost.

Within the next several years, computers will decrease in cost, and knowledge of their use will be more widespread. The computer will probably be just like the telephone. Each home will have one, and everyone will know how to use it.

the last word

One of the most significant recent developments in the computer field is timesharing. This is a process where one large computer is used to serve many individuals concurrently. Since large, powerful computers are still very expensive, they can't generally be afforded by small businesses and individuals. But occasionally, the services of the larger computer are needed. Timesharing makes it possible. Each user in a timesharing system has a small, inexpensive I/O terminal, much like the typewriter unit shown in **fig. 2,** at his home or place of business. The computer is located at some central and convenient point, and the user communicates with the computer via the telephone lines with his I/O terminal.

Timesharing places the computing power of a large computer in the hands of a number of users. Since the computer is shared, the cost is divided among them, making the price of computing very low. In fact, it's so low that in the near future all of us will probably have a timesharing I/O unit in our homes. As timesharing equipment and programming techniques are improved, their cost will no doubt become lower.

It doesn't seem as though ham radio and timesharing go together though, does it? However, there are possibilities. Why not ham-station control by timesharing? The cost would probably be cheaper than buying a complete computer. Our control program would be stored in the main computer, and because of the vast power of this larger computer it could be a very extensive and sophisticated program permitting features too complex for a smaller computer.

Perhaps we could set up a system where we could communicate with the central computer by radio rather than by the telephone lines. Timesharing by radio would be faster and cheaper, since there would be no charge for the telephone lines. Such a timesharing system could be set up for hams. Then we could all use it for ham-station control or any other application. A computer system like this, coupled with a satellite system, would put a fantastic communications capability in our hands.

It is not a question of technical feasibility, but rather of whether we should do it at all. Is it just too fantastic? We could do it right now. But maybe you don't want a computer to operate your station even if you could afford it. Then again, the prestige and convenience may be just what you are looking for. Besides, you will still operate the station yourself by controlling the computer. Or would you?

ham radio

repairing high-voltage transformers

Alf Wilson, W6NIF, 3928 Alameda Drive, San Diego, California 92103

The plate voltage transformer is pretty much taken for granted by most hams. It represents a pretty large initial investment and is expected to perform dependably with no servicing other than general cleaning and removal of accumulated dust. A transformer rated to provide power for a kilowatt rig can cost anywhere from about $40.00 to over $100.00, depending on how much horse trading you're willing to do. It is therefore not unreasonable to expect trouble-free operation of the device, providing the usual overload circuits are incorporated in its installation.

Despite good installation practice it's possible for power transformers to fail, sometimes for no apparent reason. Regardless of whether the transformer is a second-hand pole pig obtained from the power company, or a brand-new one from a well-known manufacturer, there is no absolute guarantee that something can't happen to cause a failure (Murphy's third law).

I have had two sad experiences with large power transformers. Both are commercially made units bought from reputable firms. The first failed as a result of rectifier tube flashback, and the second just plain short circuited in the secondary. Both problems occurred despite overload protection circuits. Each transformer was enclosed in a steel case, and each was wound with the secondary over the primary. One was even impregnated with tar. The tar is supposed to protect the windings from moisture penetration. (It doesn't.) The unit that failed because of rectifier flashback was easiest to repair. The other was a more involved problem and entailed complete disassembly to locate and repair the trouble; the secondary had to be completely rewound on this unit.

Both transformers appeared to be hopeless cases as far as being repairable. However, I like to operate the rig and really couldn't stand the expense of replacing a $75.00 piece of equipment. I figured I had nothing to lose anyway, so I decided to see if the transformers were worth repairing. The following paragraphs describe how these transformers, which would otherwise have ended up as expensive junk, were doctored and restored to service.

causes of transformer failure

Most power transformers fail because of high voltage insulation breakdown. It is rare for a transformer to develop an open or short circuit in the primary winding. If problems do occur in the primary (e.g., a short between turns, open circuit, short to the case), then you are faced with removing anywhere from 3000 to 4000 turns of secondary wire, usually about number-22 AWG, just to expose the primary. This, of course, is assuming the primary is wound under the secondary as is the case with many large power transformers. The primary, once you get to it, is easier to repair than the secondary since the primary consists of 300 to 400 turns of fairly large wire (about number-16 AWG).

preliminary checks

It is not necessary to remove the case for an initial check, because all you want to do at this point is try to get a clue as to where the trouble is and how serious it might be. Not much can be done with a bad transformer while still in the power supply, so the first thing to do is set it on the workbench and make some resistance measurements. First test for a short between primary and case, then between secondary and case. Next measure the resistance across the secondary and between each end of the secondary and center tap. The secondary resistance should be around 300 ohms between each end and the center tap if there is no short be-

tween turns. If the meter shows a resistance **differential** of more than 30 or 40 ohms between each end of the secondary and the center tap, then you know the problem is somewhere in the secondary.

The next step is to remove the case. Remove the bolts and save the lockwashers and nuts. You'll find upon reassembly that you must pull down the nuts on the case bolts quite tightly to minimize core hum. The lockwashers are necessary on these monsters!

The primary and secondary leads should now be exposed. These are just long enough to reach the terminals without much slack to spare, so you'll have to use reasonable care while working with them.

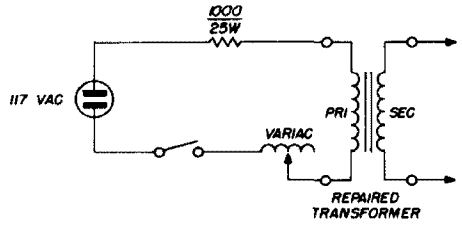


fig. 1. Circuit for "smoke testing" the repaired transformer. The 1000-ohm resistor is for current limiting in case one of the windings is still shorted.

They will stand some flexing, but it is possible to break one off where it disappears into the winding. Power transformers are pretty rugged otherwise and will withstand ordinary handling during disassembly and assembly.

Unsolder the leads from the lugs on the insulators, then dig out any tar. (Comes out in chunks—no problem.) Inspect the winding area for charred spots. These will be pretty obvious if the short was caused by rectifier flashback. If the failure was caused by some other problem then the trouble will be harder to locate, and you will have to disassemble the core.

flashback failure

My transformer caught fire when one of the rectifier tubes flashed back. For some reason I'll never know, the overload circuit just flat failed to protect the transformer. There was a charred spot where one of the secondary leads was brought up out of the transformer innards. Chances are pretty good that, with a rectifier flashback failure, the short-circuited area will be near the coil surface, and probably in the first secondary winding layer. In this type of failure, extreme heat caused by the high current surge destroys part of the insulation, allowing one or more secondary wires to touch.

Carefully peel away the insulation and clean off the carbonized material with an old toothbrush dipped in alcohol. Don't attempt to replace any of the original insulation, even if it appears to be okay. Apply at least six coats of red glyptal varnish all around the exposed wire, and work it well into the surrounding layers of insulation. It is essential that every bit of carbonized material be completely removed before applying the glyptal. Allow five full hours for each coat to dry.

Next, wrap a layer of Scotch number-33 electrical tape (rated at 10,000 volts per layer) over the exposed wires, then apply two more coats of glyptal over the tape and surrounding area. It's a good idea to slather glyptal all over the outside of the windings, again observing adequate drying time between coats. You just can't put too much glyptal on these things.

After another ohmmeter check, as described previously, the unit can be reassembled. It's really not necessary to pour melted tar around the windings. As a matter of fact, my transformer had only an inch layer of tar at the top and bottom of the case where the end bells join, so I didn't bother to replace the tar. This transformer has been working perfectly for the past five years. (By this time maybe you're getting the message that I don't have much faith in tar-impregnated power transformers. I have a strong suspicion that this stuff does more harm than good.)

The repaired transformer should now be given a "smoke test" before reinstalling it in the power supply. The simple circuit shown in **fig. 1** is used for this check. You can substitute several 115-volt light bulbs for the Variac, but they should be ar-

ranged so they can be progressively shorted to increase the primary voltage. Close the switch and slowly turn up the Variac. It should hardly be necessary to emphasize that you're dealing with upwards of 3000 volts of instant death, so use utmost care during this test. If you've done a good job of cleaning and insulating, your transformer can now be reinstalled in the power supply, and you're back in business.

digging deeper

Why should an expensive, well-made power transformer mysteriously develop a short circuit? Commercially built transformers are conservatively designed, and engineers specify high voltage insulation with large safety factors. Nevertheless, a transformer can be working fine one day, and suddenly a failure can occur in the insulation, causing a short circuit. Apparently moisture somehow penetrates into the secondary and causes a high resistance path to build up. Finally, one day the insulation just breaks down.

When you inspect the external winding area after one of these "mysterious" failures, chances are you won't see anything as obvious as in the flashback case. But don't give up—all you have to do is remove the core, then unwind the secondary to expose the short circuit. This task isn't really as formidable as it might appear. First, let's take a look at how the core is put together.

core arrangement

The core consists of E I sections of silicon steel about 0.01-inch thick. The E sections are placed over each other back-to-back, interleaved with I sections at each outer edge (**fig. 2**). You will find that the laminations have been driven quite tightly into the coil. A little strong-arm effort here will work one of the end laminations loose, then the others will come out easily. You will also find upon reassembly that it will probably be impossible to drive all the laminations back into the coil. My rewound transformer has about ten E I sections missing, but the transformer doesn't seem to know the difference.

the solenoid

As the coil is unwound, you'll have to look carefully for evidence of insulation breakdown. It's a good idea to go over each layer with a magnifying glass, because the area of failure will probably contain extremely small holes in the insulation. Be sure to count the number of turns taken from each layer and keep track of them. While unwinding my coil, I made a rough sketch of each layer, indicating the winding area and number of turns per layer. It is not worthwhile to rewind with the old wire, because no matter how careful you are, it's virtually impossible to avoid nicking the wire. Even one tiny nick will wash out all your hard work.

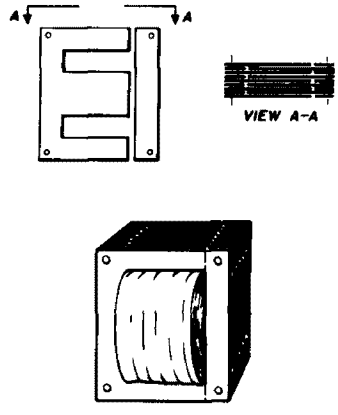


fig. 2. Core buildup and assembled transformer.

the electrostatic shield

If you have to remove the entire secondary winding, you'll probably find a layer of paper-thin copper between secondary and primary. This is a shield to prevent electrostatic coupling between the two windings. Its purpose is to keep higher order line frequency harmonics out of the d-c plate supply. This shield isn't really necessary for ham power supplies and just causes added insulation problems. If you decide to keep the electrostatic shield, make certain it is reinstalled exactly as it was originally. My recommendation is to throw the thing away.

secondary wire

If your transformer is more than ten years old it is probably wound with enam-

elled Formvar. I searched all over trying to find a source of this wire, but found it's practically nonexistent these days. I rewound my transformer secondary with HPTX-200, made by Essex Wire Corporation. This material has a high temperature insulation of some type of polymer. It costs about $2.00 per pound for number-22 AWG; it weighs about the same as Formvar, so you can determine how much you'll need after you remove the old wire.

When you've located the short circuit, splice on the new wire, wrap the splice with Scotch number-33 electrical tape and give the entire area the red glyptal treatment. You are now ready to rewind.

rewinding the secondary

I built a jig out of a piece of mop handle to hold the solenoid for the rewind job. The solenoid, which has a paper tube core, was force fitted to one end of the jig and held with a couple of wooden wedges. I turned down the other end of the jig to fit into the chuck of a large portable drill motor. I mounted the motor on the bench with the solenoid protruding about eight inches over the edge of the bench. The spool of new wire was mounted onto a U-bracket fastened to a wooden cleat. This assembly was placed on the floor beneath the solenoid.

The gears in the drill motor provide just enough drag so that the solenoid can be turned by hand during the rewind process. Maybe it's possible to apply power to the motor and rewind the coil, but I found that to control the windings you have to rotate the solenoid by hand. This takes about an hour per layer for us old folks. As you rewind, put on the same number of turns per layer that were removed—no more, no less. Don't attempt to hop up the transformer by changing the primary-to-secondary turns ratio. You are only asking for more problems, and an additional 500 volts or so on the secondary won't make that much difference to your signal anyway. (Remember, you have to **double** your transmitter power just to gain a 3-dB increase in signal strength.)

Wrap each rewound layer with Scotch number-33 electrical tape and apply two coats of red glyptal varnish on each layer. You are now ready to reassemble the core laminations.

reassembly

If you have rewound the coil as tightly as possible, you should have no trouble reinserting the core. It may be necessary to squeeze the solenoid slightly to accommodate the laminations. With one layer of tape around each layer of wire, plus the glyptal, the solenoid buildup should be very close to the original. It may be necessary to tap the last few laminations home with a mallet. Don't worry about slight bends in the laminations. When you pull the bolts down in the case, the laminations will flatten out. Now give the transformer the ohmmeter check, then the smoke test as described earlier.

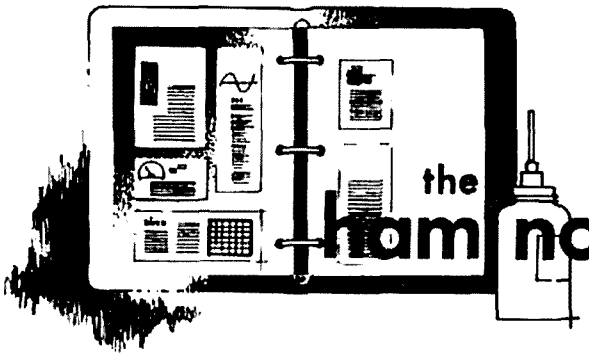
baking

An added refinement, though not necessary, is to give the repaired transformer a baked varnish treatment. For $7.50 I had this done at a local electrical shop that specializes in rewinding motors and generators. (Incidentally, while there I asked them for a quote on how much they would charge to repair my transformer. Would you believe $65.00, **not** including material?) For the baking treatment, the uncased transformer is immersed in a vacuum tank, and insulating varnish is forced into the windings. The unit is then baked for several hours. I was a little concerned as to how the baking would affect the temperature resistance of the Scotch electrical tape, but apparently it does no harm. This baking treatment is probably an improvement over tar impregnation.

concluding thoughts

You now have spent about $25.00 for material, are back on the air, and have had the privilege of being introduced to the all-but-forgotten art (at least in ham circles) of transformer winding. Hey, you old timers—kind of tugs at the heart strings, eh?

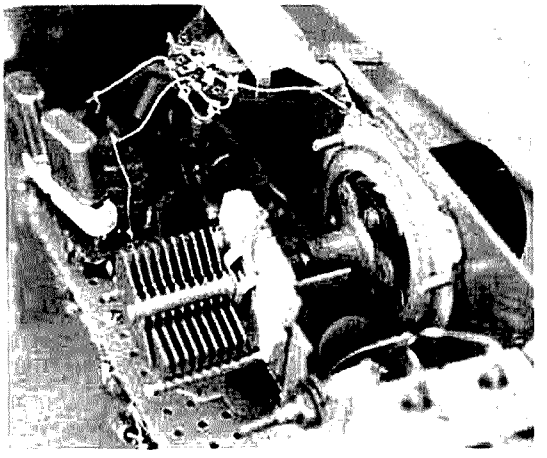
ham radio



dc crystal switching

This short article was prompted by an article by W7OE in the May issue of **ham radio.*** It's often desirable to use a single oscillator and switch crystals as a source of multiple-frequency check points. Usually a number of crystals are simply mounted around the periphery of a multiple switch. The major difficulty you run into with this approach is from the mechanical configuration that results. First, some of the rf leads are undesirably long. Second, the physical requirement for the switch shaft to stick through the front panel limits the over-all oscillator configuration.

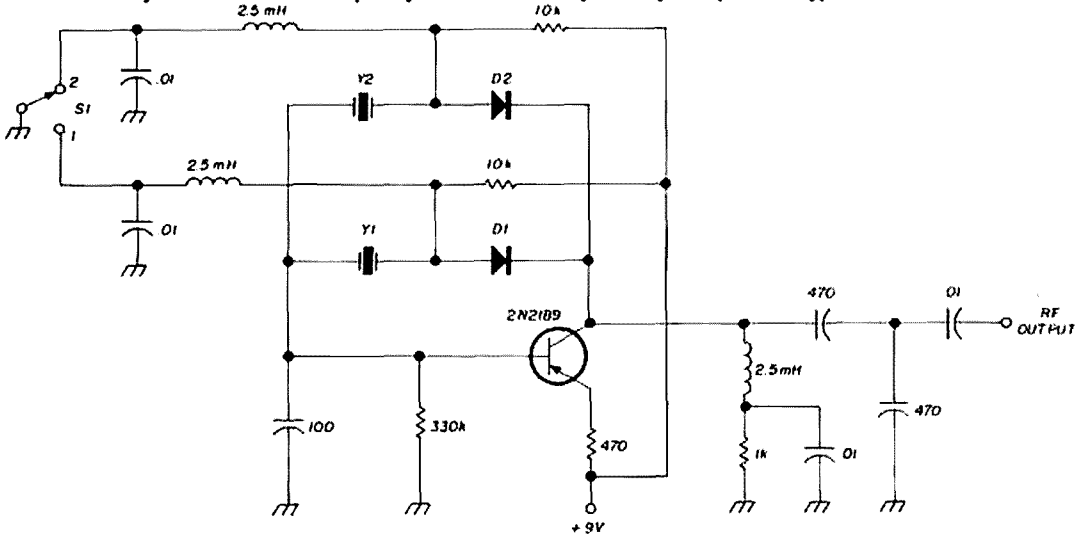
A diode switching approach allows flexibility that simplifies the layout, uses a single oscillator and handles multiple crystals. The circuit in fig. 1 shows the principle applied to the same type of oscillator used by W7OE. The beauty of this approach results from the fact that switching is accomplished with low-voltage dc.



The oscillator can be located anywhere and the dc switching wires carried to a switch located at a convenient place. Since there is no rf on these wires, their length is unimportant. The photograph shows an oscillator using two crystals for sideband selection installed in an ssb receiver. In this type of

* H. S. Pyle, W7OE, "Mini-Spotter Frequency Checker," *ham radio*, May, 1968, p. 48.

fig. 1. Diode switched crystal oscillator. With switch in position 2, diode D2 is reverse biased; D1 is forward biased and crystal Y1 controls frequency. The diodes may be any inexpensive type.



application, an additional advantage is gained from the fact that the wire only carries dc so it doesn't radiate any undesired rf into adjacent receiver circuitry. This approach could also be used for selection of crystals in ssb exciter transmitter circuitry.

R. J. Schlesinger, K6LZM

sbe linear amplifier tips

When the original owner of my new second-hand SBE SB1-LA linear boasted that he had keyed the unit at full power for over two minutes off resonance without any damage, I was skeptical; when I looked inside, I wasn't surprised at what I saw—a full kilowatt input can raise havoc in two minutes. The envelopes of the six 6JE6 tubes were discolored and bubbled, and subsequent tests indicated that only two of the tubes had any useful life left.

My first task in getting the linear back on the air was a set of new tubes. Three weeks and $22 later, I had them; however, they weren't balanced—with power applied two or three of them would glow cherry red while the rest loafed along acting as parallel capacitors. Results: overheated tubes, poor efficiency and blown line fuses. A new set of factory-balanced tubes, a look at the schematic and a few simple modifications solved the problems.

The linear was originally built with the cathodes grounded and common bias fed to all six control grids. The circuit was simple enough, but there was nothing to cut off a tube if its idling current went over the normal 50 mA. My initial modification was to install a 51-ohm, 2-watt, 5 percent resistor in each of the six cathode leads with as short leads as possible—each resistor bypassed with a .02-μF ceramic capacitor. Resetting the bias adjustment brought the idling current back up to the normal 300 mA.

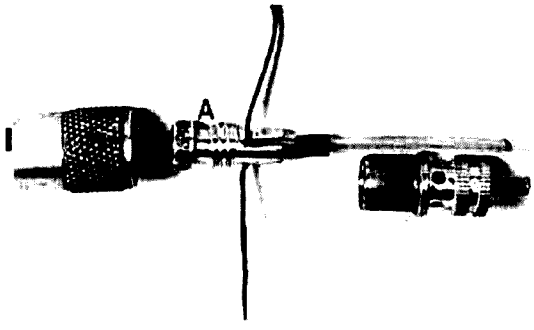
Since completing this simple modification, I have accumulated over 1000 operating hours on this set of tubes. They are not at all discolored, and recent tests indicated high emission and no internal shorts.

R. L. Wentworth, WA6DCW

coaxial cable connectors

Whenever I put a coaxial fitting on a piece of RG-48/U, I always seem to end up with a short circuit. At least one of the shield strands breaks, leaving a small piece to float around inside the connector until it shorts the whole thing out. A simple solution that works well on the high-frequency bands is shown in the photo.

All you have to do is saw two L-shaped slots in the reducing adapter (A in photo). Saw a slot down the center of the adapter almost to the threaded section with a hack saw; then saw or file a short notch at right angles to the main slot—do the same thing on both sides of the adapter. Tin the notches on each side with a hot soldering iron.



Prepare the coaxial cable by pushing it through the adapter; strip off the outer jacket, unravel the shield and twist it into two leads opposite from each other. Slip the two pieces of twisted shield down the slots and solder them into the notch. If you've done a good tinning job, this shouldn't take too much heat. Now cut off the excess pieces of braid. With a little care and minimum solder, you should have no trouble screwing the adapter into the rest of the plug.

Save a small section of the outer jacket—about a quarter inch long. This can be pushed into the reducing adapter and fills up the space around the dielectric, providing final electrical isolation. The inner dielectric is stripped off 1/16 inch in front of the adapter. When soldering the inner connector to the plug, be sure to slant it downward to keep any solder blobs from running down inside and shorting everything out.

Ted Woolner, WA1ABP

miniature receiver antenna tuner

Most modern, high quality receivers have an input impedance of 50 ohms and are designed for use with matched, transmitting-type antennas. Unfortunately, these receivers must sometimes be used with random-length wires. This often leads to inadequate performance, especially poor sensitivity and degraded image rejection.

A simple antenna tuner can be used to correct this situation. A miniature tuner which will match a short high-impedance antenna to a receiver is shown in **fig. 2.** The version shown, which was suggested by a tuner built by K4BXO, has proved useful for matching 10-to-25-foot antennas between 10 and 30 MHz. Changing the type and size of core, number of turns, and tap positions will modify the impedance matching and tuning range.

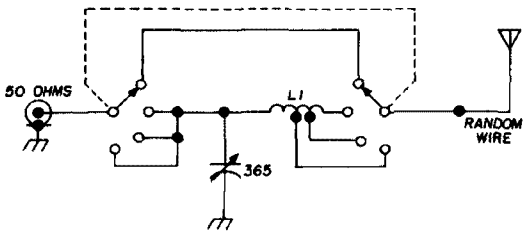


fig. 2. Receiver antenna tuner for random length antennas. For the range 10 to 30 MHz, the coil is 35 turns no. 26 on 5/8" diameter toroid, tapped at 10 and 20 turns.

The tuner can be built in a very small minibox, since few parts are required and those that are needed can be miniature ones. A toroid similar to the one I used is available from Amidon. A plastic-dielectric variable capacitor from Lafayette is excellent for C1; its capacitance is not critical. I used a miniature Alco rotary switch.

The tuner is very easy to use. Simply tune the capacitor through its range for each switch position. There should be a significant improvement in signal strength and image rejection as this is done. If no improvement is noted, the antenna length or number of active turns on the core can be changed.

Paul Franson, WA7KRE

using dipole antennas on non-harmonic frequencies

There are cases where it's desirable to use a dipole at frequencies on two bands which are not harmonically related. A good example is the phone man who wants to use a single antenna centered on 3.9 and 7.25 MHz. Since these frequencies are not harmonically related, the operator resorts to a compromise arrangement that leaves something to be desired on both frequencies.

Here's a system I use to obtain resonant operation with an off-center-fed Windom antenna. I cut the antenna as a full-wave dipole at 7.25 MHz as a starter; natural half-wave resonance is around 3.625 MHz. The resonant point is moved **up** to 3.9 MHz by the simple expedient of putting a fixed capacitor of the right value in the center. This is simple and can be used on **non-center-fed** dipoles operating on 10 and 15 meters or 15 and 20.

To compute what size capacitor you need in the center, consider the antenna as an isolated single-wire transmission line operating above a ground plane. Although the characteristic impedance of this line can be computed from the formula

$$Z_o = 138 \log \frac{4h}{d}$$

where h is the height above ground and d is the diameter of the wire, an assumed value of 600 ohms is close enough for our purpose.

The difference between a quarter wavelength at 3.625 MHz and 3.9 MHz is 4.6 feet; at 3.9 MHz, this represents 6.55 electrical degrees. The reactance of 6.55 degrees of 600-ohm line measured against a voltage node is 600 tan 6.55° or 69 ohms. Therefore, a 69-ohm capacitive reactance in series with each quarter wave will do the job; this is the same as 138 ohms in the center of the dipole. At 3.9 MHz, this is 290 pF; either 270 or 300 pF are close enough. A 1000-volt mica capacitor will be more than adequate for a full kilowatt.

To lower the frequency, the same approach may be used, but an inductance must be used as the reactive element.

Henry Keen, W2CTK



for the ham builder

Here's a bookful of circuits that can fit into the rig or accessory you're planning to build. Most of the circuits are simple ones, mainly for the Novice and Technician, but they'll suit almost anyone's operating ambition.

There are couplers, mike amps, modulators, monitors, filters, preamps, converters, receivers, transmitters, linears, a lot of accessories and some test instruments. Many tube circuits are included for hams who haven't made the move to solid-state yet, but there are plenty of transistor circuits for those who have. The projects are comparatively inexpensive to build, and parts are easy to find for them.

The author, Bert Simon, W2UUN, has put 104 separate projects into this book. He tells you only briefly how to put each one together. There is not much instruction to help with layout or mechanical arrangement; the author has assumed the builder already knows the rudiments of putting together ham gear, dressing wiring, and so on. The circuits, however, are explained where necessary, and tuning-up data is included.

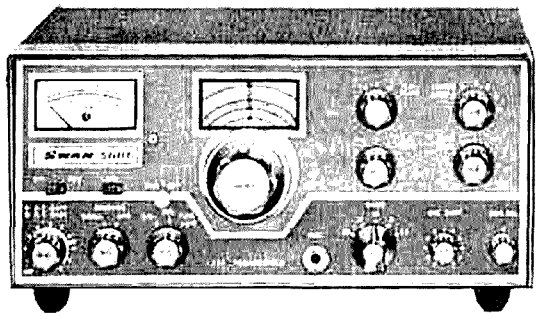
This isn't a beginner's book, but the simpler Novice circuits don't require much experience. **Ham Radio Projects for Novice and Technician;** paperback $3.95; hardcover $6.95; G/L TAB Books, Blue Ridge Summit, Pennsylvania 17214.

mobile manual for radio amateurs

In case you, as I, have and use all of the ARRL publications, you may not have noticed that a new edition (the fourth) of the **ARRL Mobile Manual** is out. The cover looks the same—though they've changed the photo—but the inside is completely different. Much has happened since 1962, when the last edition appeared, and the manual reflects it. Now the book is full of modern semiconductor equipment (and vacuum tubes where they are still needed) for all sorts of uses: receivers, transmitters and audio equipment, transceivers, ssb equipment, power supplies and test equipment. Complete sections cover noise suppression and the all-important antenna. Perhaps most interesting are the chapters devoted to portable gear, which has really come into its own lately. In addition, the text has been rewritten and reset, making an easy-to-read book you'll likely refer to often. Best of all, the price remains the same old $2.50. Anyone want an old mobile manual?

ic fundamentals and projects

There are few things in modern electronics that rival the integrated circuit's influence in circuit design and application. This new book, written by Rufus Turner, covers the technical development of the IC, its general features, types and applications. In addition to an informative discussion on how the IC is built, there are many inexpensive construction projects described, including a dc voltmeter, af-rf signal tracer, crystal-controlled frequency standard, high-gain preamplifier, and quarter-watt audio power amplifier. 75c postpaid in the U.S.A. from Allied Radio Corporation, 100 N. Western Avenue, Chicago, Illinois 60680.



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the radio amateur's vhf manual

I've owned three copies of Ed Tilton's **VHF Manual** since it came out in 1965, and have worn out two of them. When it appeared, almost every ham interested in vhf was grateful, and many who weren't became so. Now Ed, W1HDQ, **QST's** vhf editor and one of the two or three best-known ham vhf experts, has revised the manual to make it even better. The original edition appeared just when semiconductors killed receiving vacuum tubes for hams, so that version didn't have too much on transistors. The new edition more than makes up for this lack. It contains a great amount of new material as well as the invaluable information that's made the **VHF Manual** a necessity among vhf'ers. The new edition costs only $2.50, and you couldn't make a better buy. Available the American Radio Relay League, 225 Main Street, Newington, Connecticut 06111.

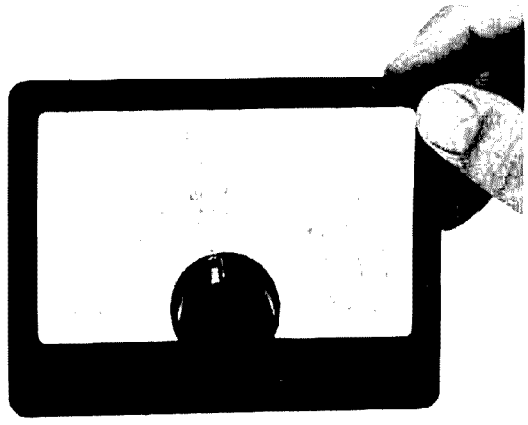
want to fix your own tv?

There's now an easy way to learn how. A brand-new low-cost TV servicing/repair course, designed for the layman, is off the press at International Correspondence Schools; it's written by **ham radio's** roving editor, Forest H. Belt.

It begins with the simplest things, like how to make adjustments and recognize tube troubles, and takes you all the way through circuit servicing. In the simplest terms you can imagine, it explains transistors and color TV as if they were no more complicated than a one-tube code oscillator.

You're already familiar with editor Belt's easy-reading articles in **ham radio;** the new six-book course is written in what he tells us is his **Easy-Read** format. All of the books are profusely illustrated, and the two that go into color-TV servicing use full-color pictures.

The whole course is written so plainly that anyone who can read can learn to repair a TV set. For a ham, who already knows electronic basics, the course is duck



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soup. The section on servicing logic—called 1-2-3-4 troubleshooting—lays out a technique that could be applied to ham gear or any kind of electronics. International Correspondence Schools, Scranton, Pennsylvania 18515.

silicon power transistors

Motorola has announced a new line of 4-ampere power transistors that combine high current, top efficiency and power-handling capability with economy prices. These new devices can be used in npn/pnp pairs to gain all the advantages of direct-coupled complementary symmetry circuitry.

The exclusive Thermopad ™ construction results in low thermal resistance and minimum derating in all chassis mounting applications. In addition, the compact package is simple to mount in virtually any place or position. The pnp devices of this new series are the 2N5190, -91, and -92; there pnp complements are the 2N5193, -94 and -95, respectively. For more information, write to Motorola Semiconductor Products, Inc., P. O. Box 955, Phoenix, Arizona 85001.

handbook of semiconductor circuits

If you're looking for ready-made transistor circuits, this new handbook contains 124 examples of standard transistor circuits, complete with operational data for amplifiers, oscillators, logic and switching circuits and power supplies. All in all, a comprehensive source of well-designed examples of contemporary circuits. Since a design philosophy section is included with each group of circuits, this volume won't go out of date as new transistor types are introduced. Each circuit description includes any unique design or operational data along with a schematic diagram; very helpful if you're working on transistor projects in your shop—must reading for laboratory and engineering technicians involved in transistor design work. $7.95 from TAB Books, Blue Ridge Summit, Pennsylvania 17214.

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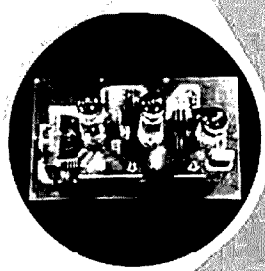
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ham radio magazine

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staff

editor
James R. Fisk, W1DTY

roving editor
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vhf editor
Nicholas D. Skeer, K1PSR

associate editors
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Alfred Wilson, W6NIF
James A. Harvey, WA6IAK

art director
Jean Frey

publisher
T. H. Tenney, Jr. W1NLB

offices

Greenville, New Hampshire 03048
Telephone: 603-878-1441

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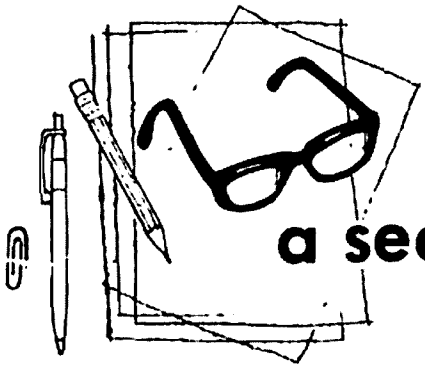
58 **flea-power solid-state transmitter**
Robert M. Brown, K2ZSQ

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Al M. Brogdon, K3KMO

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a second look

How often do we hear that ham radio is becoming an old man's hobby? We aren't attracting youngsters into our ranks at the rate that we should; the total count in most lower license categories is actually dropping. We need young people in amateur radio for many different reasons. Their enthusiasm and inquisitiveness have been responsible for much in the past and activities which tend to fall to our younger colleagues have been very important in continuing government support. Our hobby is well known as a technical training ground and obviously the younger people gain the most. If amateur training can start future engineers and technicians on their way, we are all gainers. In times of emergency the backlog of technically trained radio operators has always been an important national asset, but if this pool is all over 40 it is of far less value.

We need people who are willing to move quickly into new ground. Recent years have seen things such as the OSCAR program, Project Moonray and moonbounce. Who knows where we may head next? We'll need to be fast and adaptable, however, if we're to hold our vhf and uhf allocations against the communications technology of the 70's and 80's; and we had better have the manpower to do it!

But I think there is something of even greater importance here: youth has that wonderful energy which can accomplish so much when properly directed. We can always use this boost. New questioning of old ideas? It won't hurt us. The worthwhile traditions will survive while the outmoded will be replaced with more timely approaches. Remember that our hobby was started by young men who were willing to try something new.

Many factors have slowed the influx of newcomers to our ranks. The ease of getting on the air via CB has taken recruits; our recent political dissentions have surely cost us members. However I expect that we have been the biggest problem ourselves. I have been dumbfounded to meet would-be hams who have been unable to find any local help in obtaining a ticket and getting on the air. Some have actually been turned away by local hams and clubs. Unbelievable, but unfortunately, true.

What can we do? Lead the way. Invite those young neighbors over and show them what it's all about; explain your equipment; let them have a shot at the mike. Let them know that the welcome mat is always out and that they are really wanted in amateur radio. If you have a local club, talk up a recruiting and training program. You'll be the winner as you get some of the most enthusiastic members you ever had.

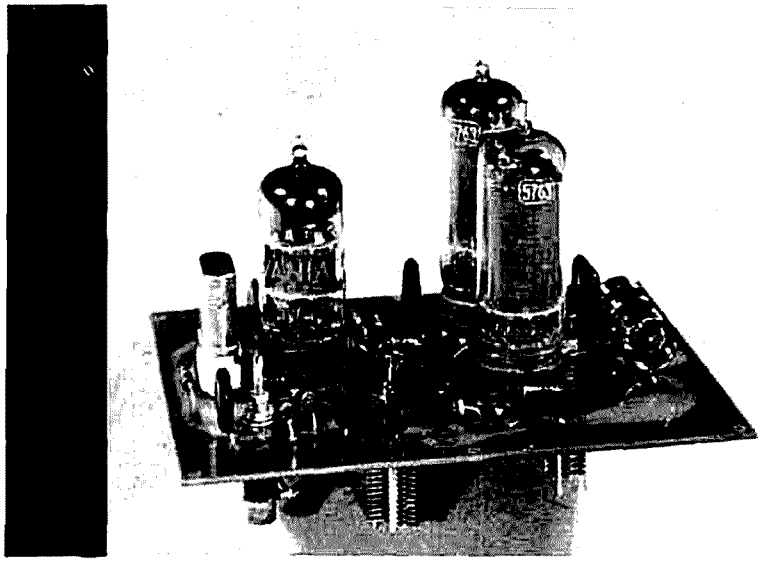


Left to right: WA6ONZ, WN7LIX and KØTFL.

The picture shows Buzz MacDonald, WN7LIX, being awarded a Mosley Tri-band beam by Jack Mosley, KØTFL, of Mosley Electronics, Inc., and a Tri-Ex tower by Clyde Blyleven, WA6ONZ, from the Tri-Ex Tower Company. What for? Because WN7LIX was the youngest licensed ham (age 9) at the SAROC convention in Las Vegas last January.

We can't all give away beams and towers, but some of your old equipment mixed with a bit of patience and encouragement could go a long way towards giving our hobby a real boost.

Skip Tenney, W2NLB
Publisher



heterodyne transmitting mixers for six and two meters

All you need is an 80- to 10-meter exciter to participate in the fun on the vhf bands with these transmitter converters

D. W. Bramer, K2ISP, 45 Thayer Road, Fairport, New York 14450

The two units to be described were designed primarily to get the average high-frequency (80-10 meter) enthusiast off to a good start on the vhf bands with minimal cost and effort. They also provide interesting and straightforward construction projects for those amateurs who, despite all the easy approaches, respect ham radio as a technical avocation. Today's ham who knows his way around is sure to get more out of his hobby than the purchaser of commercially available boxes. Collect the necessary parts, put one or more of these units together with your own hands, adjust them properly, and you're off with tremendous satisfaction to explore the exciting world above 50 MHz.

Both units are of readily duplicated printed wiring board construction. Each is capable of producing about one watt of ssb or CW output, either directly to an antenna or as input to a higher-power linear amplifier.

The 50-MHz satisfies the objective of using almost any existing 80-10 meter transmitter as a basic rf source to generate signals at vhf, while retaining the operating convenience of the low-frequency equipment, such as vfo, upper or lower sideband selection, vox, PTT and break-in CW. It does so by accepting 7-MHz input and heterodyning it to the six-meter band.

Seven MHz input was selected because there are numerous single- and tri-band transceivers available, especially on the used market, at low cost. Also, this band represents a good compromise between receiver stability and freedom from undesirable transmitting-mixer output products.

Unlike the six-meter unit, the two-meter mixer accepts a higher input frequency of 50 MHz. It will, therefore, produce two-meter output when driven by six-meter input. Hence, the six-meter mixer unit may be used directly to excite the two-meter mixer. You might then assemble the six-meter unit first and consider the two-meter unit project later on.

station layout

Although many amateurs tend to concentrate on a single band and/or mode of operation, there's much to be said for station versatility. With some foresight in station planning, the hf/vhf enthusiast can incorporate several bands and modes with little difficulty, thus taking full advantage of existing 80-10 meter gear and newly added equipment. **Fig. 1** outlines a typical combination hf/vhf station, showing major rf paths and switching. While this block diagram is fundamental, it will aid the newcomer in understanding just what vhf heterodyne transmitting mixers are and how they may be effectively integrated with existing equipment. It may even stimulate some thinking on the part of "mister experienced" toward more sophisticated or improved station arrangements. Muting bias and relay control lines are not shown, but are discussed briefly later on.

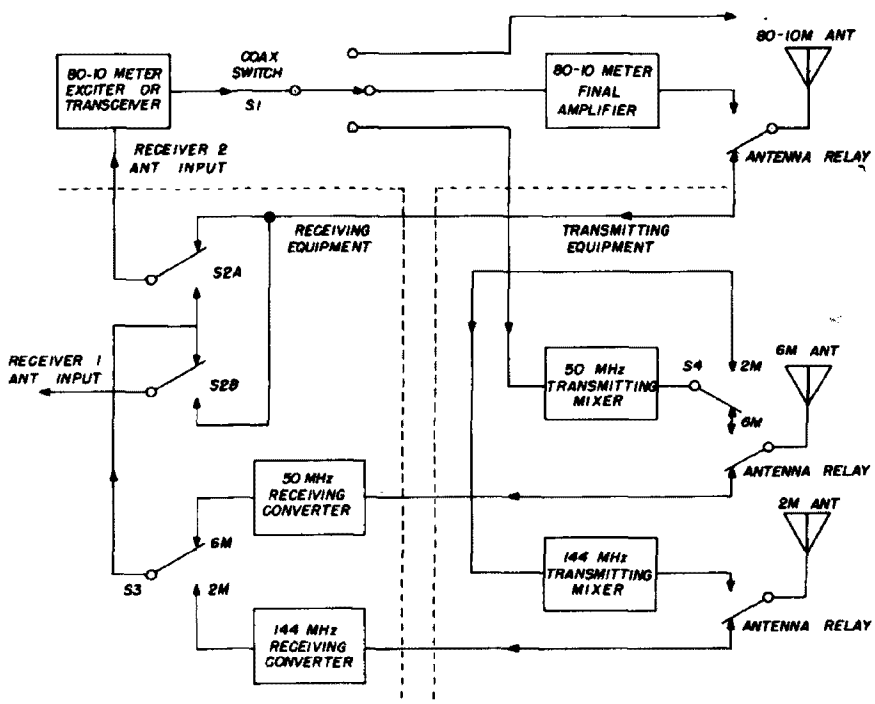
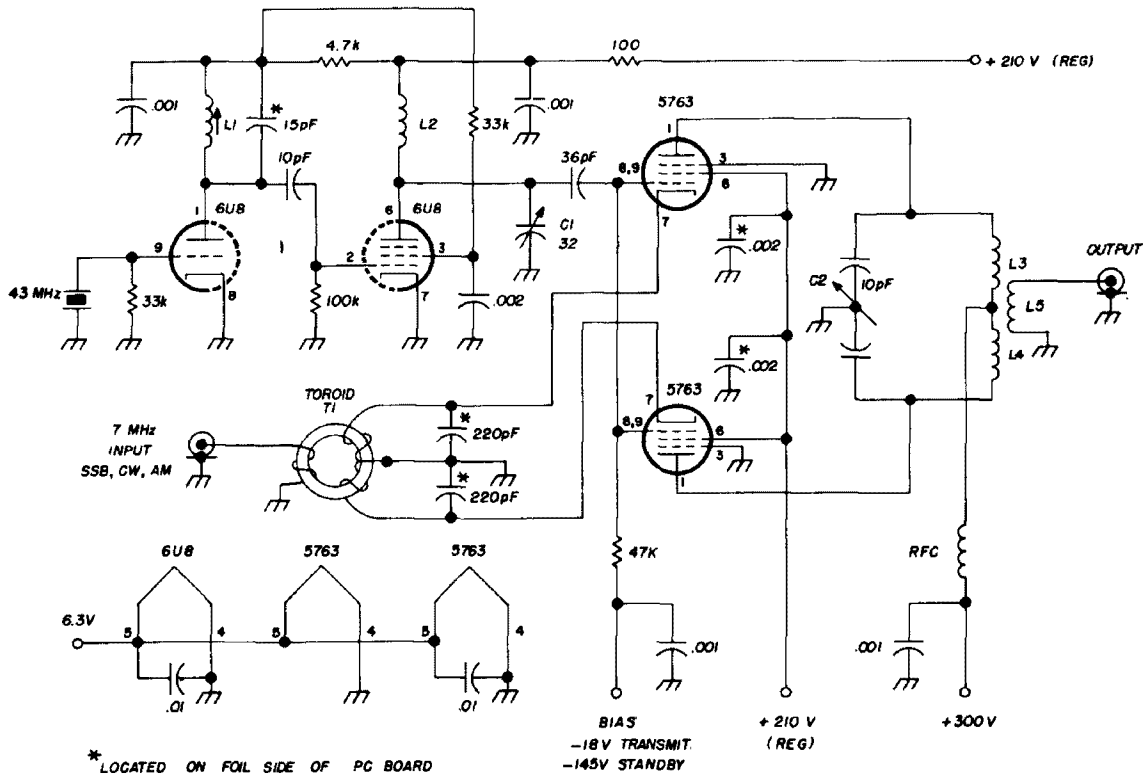


fig. 1. Typical hf/vhf rf switching system. Note that with the vhf converter output switched to receiver 1, the high-frequency antenna is automatically switched to the transceiver (receiver 2) by S2.

50-MHz circuit design

The six-meter mixer unit, **fig. 2,** starts out with a conventional triode oscillator, followed by a tetrode buffer amplifier; both stages use halves of a single 6U8. Oscillator output at 43 MHz is lightly coupled to the buffer, which reduces pulling effects. The buffer operates straight through on 43 MHz with its output capacitively coupled to mixer grids.

Two 5763 pentodes operate as a push-pull balanced mixer, heterodyning 43-MHz oscillator injection with 7-MHz input for an output at 50 MHz. Note that the 5763 control grids are driven in parallel, while the plates are in push-pull. This cancels the undesirable 43-MHz output product, because in-phase plate currents are produced at each end of the balanced tank circuit. Output at 43 MHz would otherwise



*LOCATED ON FOIL SIDE OF PC BOARD

C1 32-pF variable (E. F. Johnson 160-110)

C2 10-pF butterfly (E. F. Johnson 160-211)

L1 9 turns no. 24 AWG, close wound on 1/4" diameter slug-tuned form

L2 9 turns no. 20 AWG, air wound, 1/4" diameter, 3/8" long

L3, L4 7 turns no. 20 AWG, air wound, 1/2" diameter, 9/32" long

L5 4 turns no. 20 AWG, air wound, 1/2" diameter, 1/8" long

RFC Ohmite Z-50

Xtal 43-MHz third overtone (International Crystal type FA5)

T1 toroidal transformer (see fig. 3)

fig. 2. Schematic diagram of the six-meter transmitting mixer.

be prevalent due to its proximity to the output passband of the 50-MHz tuned circuit.

Forty-meter excitation is applied in push-pull to the 5763 cathodes. A miniature toroidal transformer with center-tapped secondary derives push-pull cathode drive from the unbalanced link-coupled input. Shunt bypass capacitors of 220 pF from each 5763 cathode provide a low-reactance ground at both 43 and 50 MHz. These present a high reactance at 7-MHz and, therefore, have negligible effect on the broadband toroid transformer input. The 50-MHz output is link coupled from the push-pull tank coil.

High-voltage supply requirements are +210 V regulated at about 20 mA and +300 V at 50 mA maximum. The +210 V may be obtained from two series-connected OB2 regulator tubes. Filament voltage is 6.3 Vac at 1.95 amperes. Under transmit conditions, mixer control-grid bias must be held at −18 V and switched to −145 V for muting during receive or vox dropout periods.

50-MHz mixer construction

An etched circuit board measuring 4 by 5 inches serves as a chassis for the six-meter unit.* It should be made from 1/16-

* Complete circuit boards etched, solder-tinned, and drilled are available from Stafford Electronics, 427 So. Benbow Road, Greensboro, North Carolina. Order: 6M-11 (six meters) and 2M-11 (two meters); prices: $4.10 and $5.75 each respectively.

inch thick epoxy copperclad stock. Three 0.104-inch-diameter holes are drilled along each longer board edge for mounting purposes. Drill two holes of the same diameter for the crystal socket. Tube socket holes are 0.067-inch diameter. The two miniature variable capacitors require 1/4-inch holes, while the slug-tuned coil-form hole is 11/64-inch diameter. All remaining holes may be drilled 0.040-inch diameter for coil, resistor, and capacitor leads.

Wire size and turns data for all coils are listed; L2, L3, L4 and L5 must be tight wound, using a mandrel of slightly less diameter than the coil size. A drill bit is handy for this purpose. Coils are then assembled to the PC board and coated with clear nail polish, or other dope, to retain mechanical stability. Details for the toroidal input transformer are given in **fig. 3.** It, too, is held in position on the PC board by its own leads.

Miniature 50-ohm coax such as RG-174/U is recommended for rf input and output connections to reduce stress on PC board copper strips. While this coax has adequate power capability, it should be used only for very short runs to avoid loss at 50 MHz.

Five capacitors, with short leads, are soldered to the foil side of the board. These consist of two 0.002-μF 5763 screen bypass capacitors, two 220-pF 5763 cathode bypass capacitors, and one 15-pF oscillator plate tank capacitor.

Tube sockets are all 9 pin and should be the low-loss mica-filled type designed for 1/16-inch thick PC board mounting, such as Cinch-Jones type 9 PC or equivalent. Component location may be determined quite easily by examining the photographs and the schematic. The entire PC board should be cleaned of rosin after soldering.

Full size printed-circuit-board layout for the six-meter converter.



adjustment and operation of the 50-MHz mixer

Using a grid-dip oscillator, preset the oscillator coil slug and buffer stage plate variable capacitor for a resonance of 43 MHz. Preset the mixer plate butterfly variable for an output tank resonance of 50 MHz. Insert a 43-MHz third-overtone crystal, apply 6.3 Vac filament power, and temporarily connect +210 V to oscillator and buffer amplifier stages through the 100-ohm decoupling resistor. Quickly adjust L1 for oscillation and L2 for maximum output, as indicated on the grid-dip meter, which has been switched to its wavemeter operating mode. Optimize crystal oscillator operation by turning the L1 slug until oscillation ceases, then turn in the opposite direction until the circuit just snaps into oscillation.

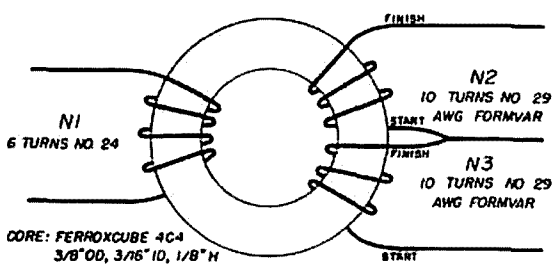


fig. 3. Toroidal input transformer for the 6-meter transmitting converter. Both the primary and secondary windings are wound around the full circumference of the core. Windings N2 and N3 are wound in the same direction. Start of N2 is connected to finish of N3 to form center tap. Toroid core is Ferroxcube 4C4, 3/8" OD x 3/16" ID x 1/8" high.

Now connect the —18-volt bias, +210-volt screen, and +300-volt plate supplies to the 5763 mixer stage. A small plate current should be evident, measured in series with the +300 volt supply line. Readjust buffer tank L2 for maximum mixer plate current. This should read about 15 mA. Check system stability by removing the 43-MHz crystal. Mixer plate current should drop to near zero. Reinserting the crystal should promptly restore the 15-mA current.

When the system responds properly to the setup procedure, you're ready to produce 50-MHz output from 7-MHz input. If more than one watt of drive is available (and it usually is), a pad must be connected between the low band transmitter and toroidal transformer input link. In my case, an SB-101 was used as an exciter. This produced about 100 watts output on 40 meters and is probably typical of most exciters. A suitable pad for this power level is shown in **fig. 4.**

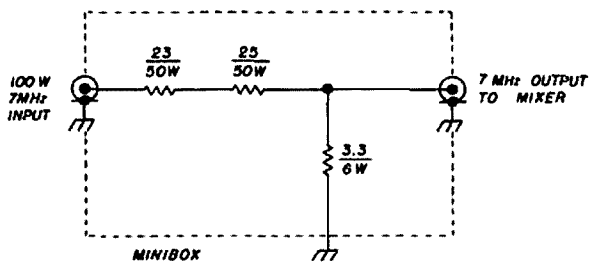
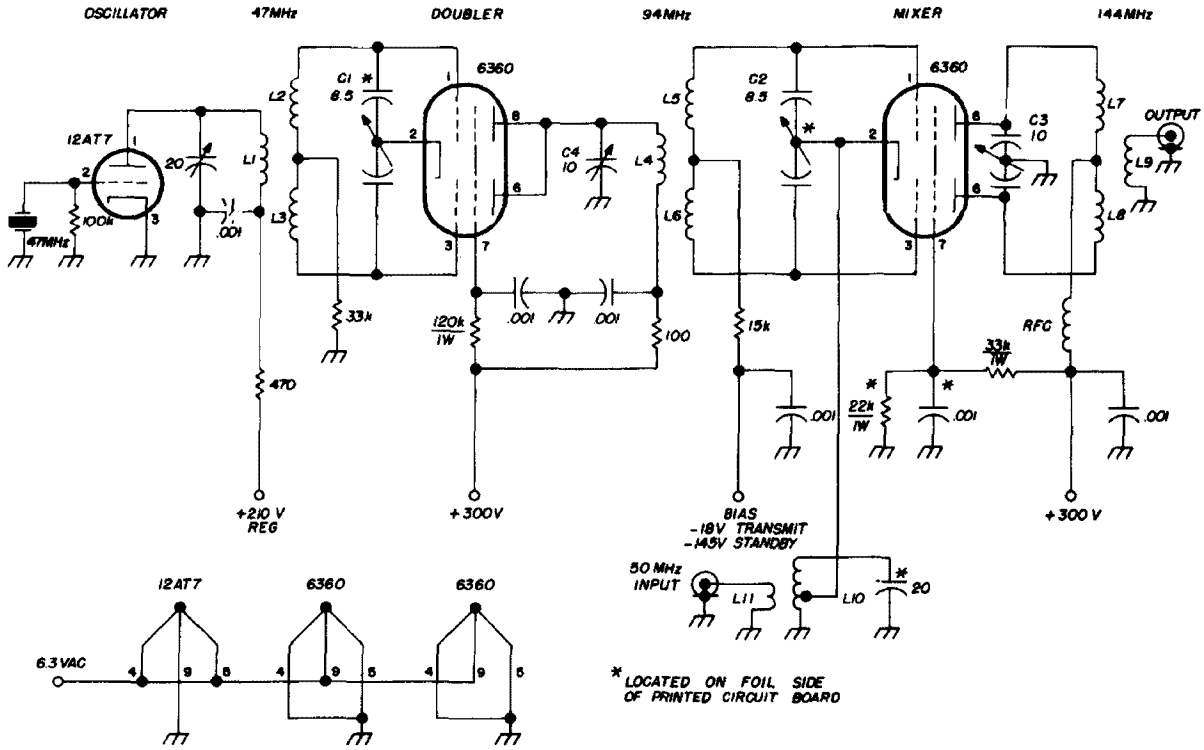


fig. 4. A 100-watt pad for 7-MHz mixer drive. 25-ohm resistors are Ohmite type 210-50; 3.3-ohm resistor is three 10-ohm, 2-watt resistors in parallel.

With all supply voltages applied and the 40-meter source connected, carefully increase 7-MHz drive until the wavemeter, held close to the push-pull mixer plate tank, indicates 50-MHz output. Peak the butterfly variable capacitor for maximum output. Again, check stability by reducing 7-MHz drive. The 50-MHz output should correspondingly decrease.

Connect a dummy load to the output link. With the load applied, tune the butterfly variable for maximum output, as indicated on an in-line forward power meter. Increase 7-MHz drive until no further increase in 50-MHz output occurs. Peak excitation on ssb should be limited to a level slightly below the above condition to resolve maximum output consistent with maximum linearity.

Connect your antenna, tune in a station, and establish contact. You'll be amazed at the results with only one watt output.



C1, C2	miniature 8.5-pF butterfly (E. F. Johnson 189-253-5)
C3	10-pF butterfly (E. F. Johnson 160-211)
L1	8 turns no. 20 AWG Formvar, air wound, 3/8" diameter, tightly wound (note position relative to L8 and L9 in photograph)
L2, L3	11 turns no. 20 AWG Formvar, air wound, 3/8" diameter, tightly wound
L4, L5, L6	5 turns no. 16 AWG tinned copper wire, air wound, 3/8" diameter; turns spaced one-half wire diameter
L7, L8	4 turns no. 16 AWG tinned copper wire, air wound, 3/8" diameter; turns spaced 1 wire diameter
L9	2 turns no. 16 AWG tinned copper wire, air wound, 3/8" diameter; spaced 1 wire diameter
L10	9 turns no. 20 AWG Formvar, close wound on 1/4" diameter slug-tuned form, tapped 3 turns from cold end
L11	2 turns no. 20 AWG on cold end of L10
RFC	Ohmite Z-144
Xtal	47-MHz third overtone (International Crystal type FA5)

fig. 5. Schematic diagram of the two-meter transmitting converter. Printed circuit boards are available for $5.75.

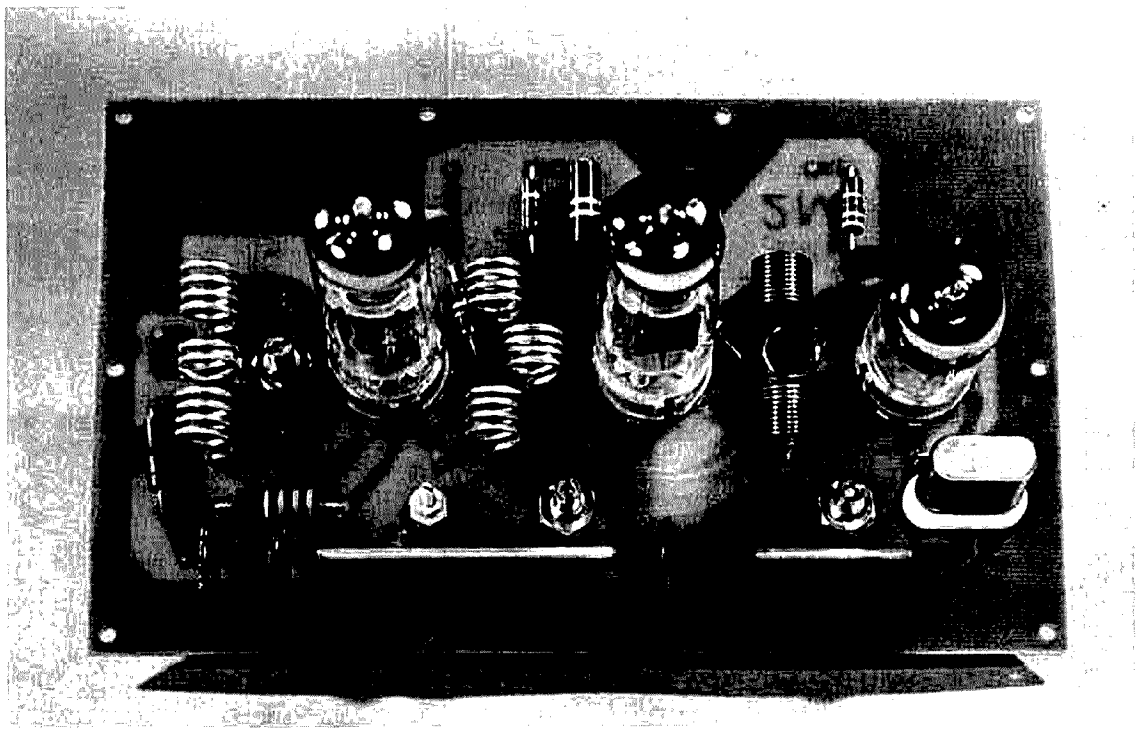
144-MHz circuit design

The two-meter mixer unit uses one-half of a 12AT7 as a conventional triode oscillator, using a 47-MHz third-overtone crystal (fig. 5). Oscillator outout is inductively coupled to the push-pull tuned grid circuit of a 6360 push-push doubler stage. With push-pull connected grids and parallel plate circuit, it doubles efficiently to 94 MHz, while canceling the 141-MHz third-harmonic product. Additional 144-MHz tuned circuits are therefore not required (as in other circuits) to reject 141 MHz from the two-meter output passband. The 94-MHz mixer injection is inductively coupled to push-pull grids of another 6360 operating as a balanced mixer.

In contrast to the six meter unit, both grids and plates may be push-pull connected, because the 94-MHz injection signal is sufficiently removed from the desired 144-MHz output frequency. The 50-MHz excitation is applied through simple link coupling to a single-ended cathode input. Again, a push-push condition exists,

which cancels both 50 MHz and third-overtone 150-MHz products from the mixer output. A link is provided for coupling 144-MHz output to antenna or linear amplifier.

Power supply requirements are similar to those of the six-meter unit, being +210 V regulated at only 8 mA, +300 V at about



110 mA maximum, and 6.3 Vac at 1.94 amperes. The same —18 V/—145 V bias is also used for transmit and muting. As a practical matter, the power supply should be capable of supplying both the six- and two-meter units.

144-MHz mixer construction

The etched circuit board for the 2-meter unit is slightly larger than that for the six-meter model, measuring 4 inches wide by 6-5/8 inches long. Again, 1/16-inch thick epoxy glass, copperclad material is recommended. Ten 0.104-inch diameter holes are drilled along the edges for mounting parts, four on each side and one on each end. The crystal socket requires two additional 0.104-inch holes. Tube-socket pin holes are 0.067-inch diameter. Holes for the three miniature variables are 1/4-inch diameter. Drill an 11/64-inch diameter hole for the slug-tuned coil form, centering it within the ring designated on artwork. Mounting holes for free-standing coils L4 through 19 are 0.066-inch diameter, while all remaining holes for L1, L2 and L3, choke, resistors, capacitors, and filament jumper wires are 0.040-inch diameter.

Coils L4 through L9 are air wound, using wire diameter spacing between turns. L1, L2 and L3 are tight wound and secured with coil dope. Oscillator plate coil L1 is mounted vertically between L2 and L3 to optimize coupling. Several components, indicated with an asterisk on the schematic, are mounted on the foil side of the circuit board. PC type 8.5-pF variables are soldered directly to appropriate foil areas, using care to maintain physical balance at the push-pull grid circuits. The 20-pF capacitor is soldered directly across the slug-tuned coil-form end terminals.

A short length of RG-174/U 50-ohm coax

may be used for the 50-MHz input connection, but larger type RG-58/U is recommended for the output connection. Here again, avoid long runs, especially at 144 MHz. If long runs are required, step up to larger coax, using only a few inches of smaller cable for board terminations.

Examination of the photographs and the schematic will aid construction. Clean the entire board after assembly.

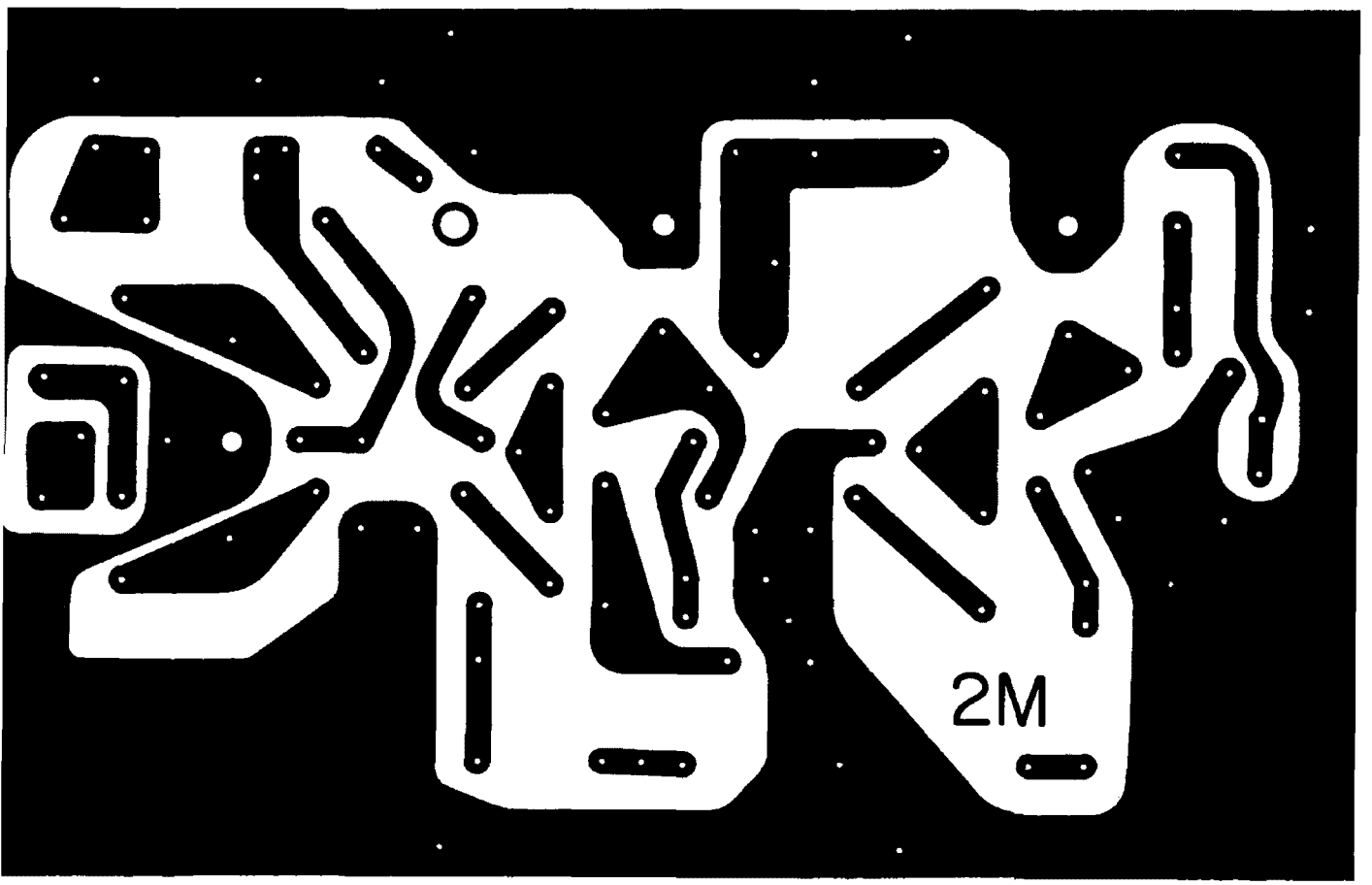
adjustment and operation of the 144-MHz mixer

Preset the 12AT7 oscillator plate and 6360 doubler grid tank circuits to 47-MHz, using a grid-dip oscillator. These circuits are rather tightly coupled due to the proximity of their coils. Resonance of each circuit can be more readily recognized by **temporarily** loading the tank circuit **not under adjustment** with a 1000-ohm resistor. The unloaded circuit will exhibit a higher Q and therefore resolve a more pronounced meter indication. The 6360 doubler plate and mixer grid tank circuits may be preset to 94 MHz, using the same procedure outlined above. Next, preset the mixer plate/output tank to about 145 MHz and the slug-tuned input coil to 50 MHz.

Insert a 47-MHz third-overtone crystal, apply 6.3 Vac filament power, and connect +210 V to the oscillator stage through the 470-ohm decoupling resistor. Adjust the variable capacitor quickly for 47-MHz oscillation, as indicated on a wavemeter. Next, apply +300 V to the 6360 doubler plates and screen grids. With the wavemeter set for 94 MHz, adjust the doubler plate variable capacitor for maximum indication. Readjust the doubler grid butterfly capacitor (on foil side of board) for maximum 94-MHz plate-tank output.

Temporarily connect a 4.7k-ohm resistor between the cold end of the 15k-ohm mixer

Full size printed-circuit-board layout for the two-meter converter.



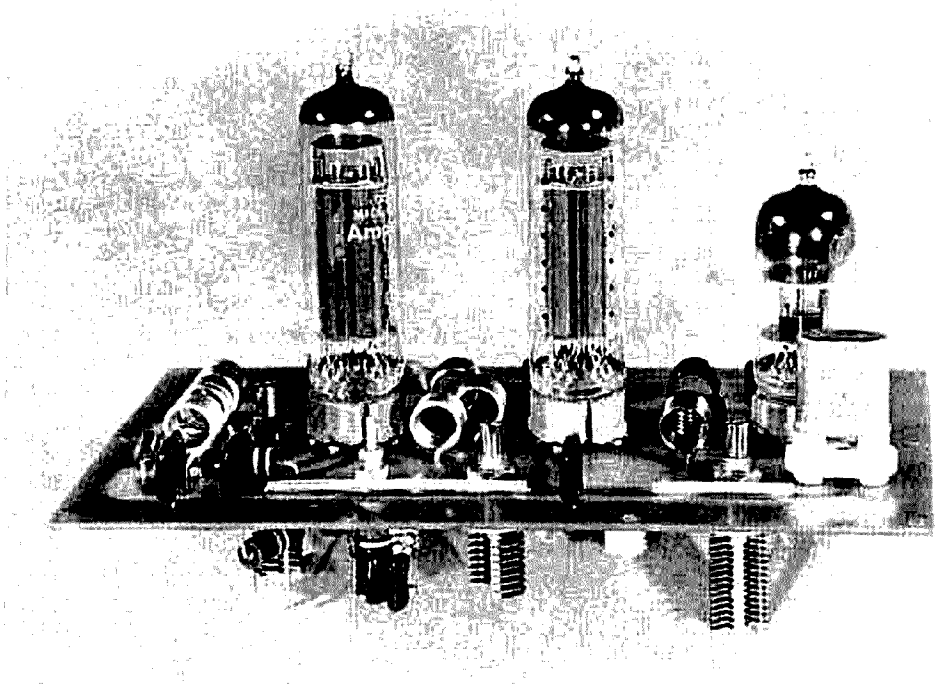
grid resistor and ground. Do **not** connect the +300 V plate/screen supply to the 6360 mixer stage at this time. With a vtvm connected between ground and the junction of the 15k- and 4.7k-ohm resistors, reconnect the +210 and +300 V supply to the oscillator and doubler. Carefully peak the doubler grid, doubler plate, and mixer grid variables for maximum negative voltage indication on the vtvm. Optimize crystal oscillator stage operation by adjusting its plate tank variable until the circuit just snaps into oscillation as with the six-meter unit.

Remove the temporary 4.7k-ohm resistor, and connect all supply voltages, including —18-volt bias. The 6360 mixer combined plate and screen current should be 40 to 45 mA. If this current level is not achieved, carefully repeak all preceding tuned circuits for maximum mixer plate current. Check circuit stability by removing the 47-MHz crystal. Mixer plate/screen current should drop immediately to 4 or 5 mA. Reinserting the crystal should promptly restore the 40-to 45-mA level.

A Heath Sixer (HW-29) makes an ideal 50-MHz source for driving the two-meter mixer board, especially during initial setup and trial operation. (In my case, the output from a Gonset 6-meter Communicator was padded down to produce about one watt of drive.)

With all supply voltages connected, apply 50-MHz drive, and observe 144-MHz output with a wavemeter held close to the 6360 mixer push-pull plate coil. Peak the butterfly variable capacitor for maximum 144-MHz output. The 144-MHz output should disappear when 50-MHz drive is removed.

Connect a dummy antenna to the output link. With 50-MHz drive applied, tune the stage for maximum output, using an in-line forward power meter. The two-meter unit will produce a reasonably good a-m signal when driven directly by a Heath Sixer or equivalent. Connect the an-



Two-meter transmitting converter.

tenna, call a local friend on the phone, and arrange to have him listen for the signal. If you have previously constructed the six-meter mixer unit, substitute its output for the a-m drive and get a report on two-meter ssb. You'll be pleased with the results.

notes

The value of adding a linear power amplifier stage between mixer output and antenna will soon be appreciated after initial experimentation with only one watt output. There are several constructional routes you might take. I assembled a complete six- and two-meter unit, including a 100-watt linear amplifier stage for each band, on a single chassis. The chassis was then enclosed in a modern-styled cabinet that matched the color schemes of existing station equipment.

As an alternate solution, you could combine a single mixer PC board with a linear output stage on a chassis. An 829B makes an excellent amplifier when operated class AB_1 at +800 V plate, +210 V (regulated) screen, and about —22-V grid bias. Many simple 829B amplifier circuits are in vhf handbooks and magazine articles. A standard 3x5x10-inch chassis offers ample space for mounting the six-meter mixer and 829 output stage, while a 3x5x13-inch chassis would nicely support a two-meter version.

If eventual construction of both six- and two-meter mixer units is contemplated, a switch should be included in the six-meter model for shifting 50-MHz output either to the input of a linear amplifier, or to the two-meter mixer unit input. More elaborate switching could be used to apply only 6.3-V filament power to sections of the system in immediate use.

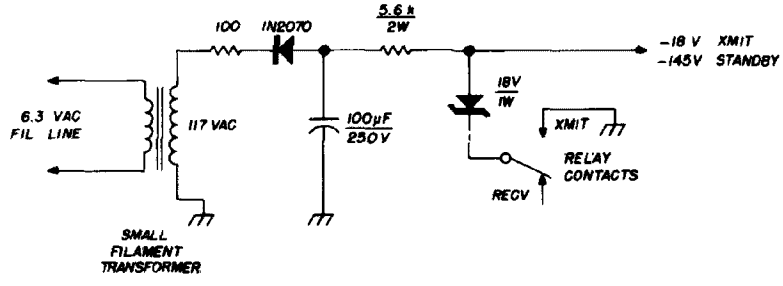
A simple bias supply (**fig. 6**) may be provided by connecting a small 6.3-V filament transformer backwards to the filament line and rectifying the 117-Vac output of the primary winding.

Relay contacts may be spare contacts on the low-band transmitter. When in nonenergized "receive" position, the zener is ungrounded and has no control, allowing the bias line to assume full —145-volt potential for muting. In "transmit" position, the relay contacts switch the zener circuit to ground, establishing proper —18-V operating bias level.

Although I chose 7-MHz input for the six-meter heterodyne mixer and 50-MHz input for the two-meter version, other frequency combinations are possible. It's only necessary to change injection oscillator crystal frequency and re-resonate oscillator and buffer (or doubler) tank circuits appropriately. As an example, consider 10- rather than 40-meter input for the 50-MHz mixer. The 28-MHz band has long been a favorite of vhf'ers, because it offers wider frequency coverage and reduced receiver feedthrough. In this instance, the crystal would be changed to 50-minus 28-MHz, or 22 MHz.* Oscillator/buffer stage plate circuits would then be tuned to the same frequency.

ham radio

fig. 6. Bias supply for the two transmitting converters.



* Or, 50 MHz plus 28 MHz, 78 MHz, for better suppression of spurious signals. **Editor**

a programmable repeater identifier

You can ragchew all day long and let this solid-state system automatically identify your station

John Connors, W6AYZ, 920 Sir Francis Avenue, Capitola, California 95010

During the past year a number of people have offered ideas on the design of a CW code generator using transistors and logic circuits. There were two major drawbacks to the systems proposed. First, it was not possible to program the CW circuit to provide a variable format that could be changed easily; second, the construction of such a device required the builder to know how to design diode matrices or other more sophisticated circuits.

The devices described here eliminate all these undesirable characteristics and allow anyone to construct his own solid-state repeater identifier or CW code generator. I'll describe the complete repeater identifier, including its electronic clock and associated control circuit. Those who are interested only in the CW code generation technique may skip down to the part titled **memory planes.**

basic chassis

This is the main chassis of the unit shown in the photographs and in **fig. 1.** It contains the power supplies, card plugs, external connectors and interface circuits for interconnecting the cards. A 12-volt lamp and a push-to-test button are on the front panel. Jacks J1 and J2 are eight-pin octals that connect the repeater transmitter and command decoder.[1] K1 is a small 10k ohm sealed relay.

memory planes

These are Vector cards that store the morse code letters. They consist of numerous one-shots in series, each one triggering the next in line as shown in **fig. 2.** A diode attached to every other collector forms a keying line that turns the tone oscillator on and off. The amount of time the one-shot is off is determined by the time constant at its base. By varying the value of R, as shown in the CW generator drawing, you can select various band lengths.[2] The 1k ohm series resistors keep the gain of each stage down to prevent oscillation. The transistors should be npn silicons with betas of around fifty. While this sounds like a lot of transistors, bear in mind that the longest calls, such as 'DE WA6XXX,' require only fifty-two transistors. In quantities the prices are quite reasonable.

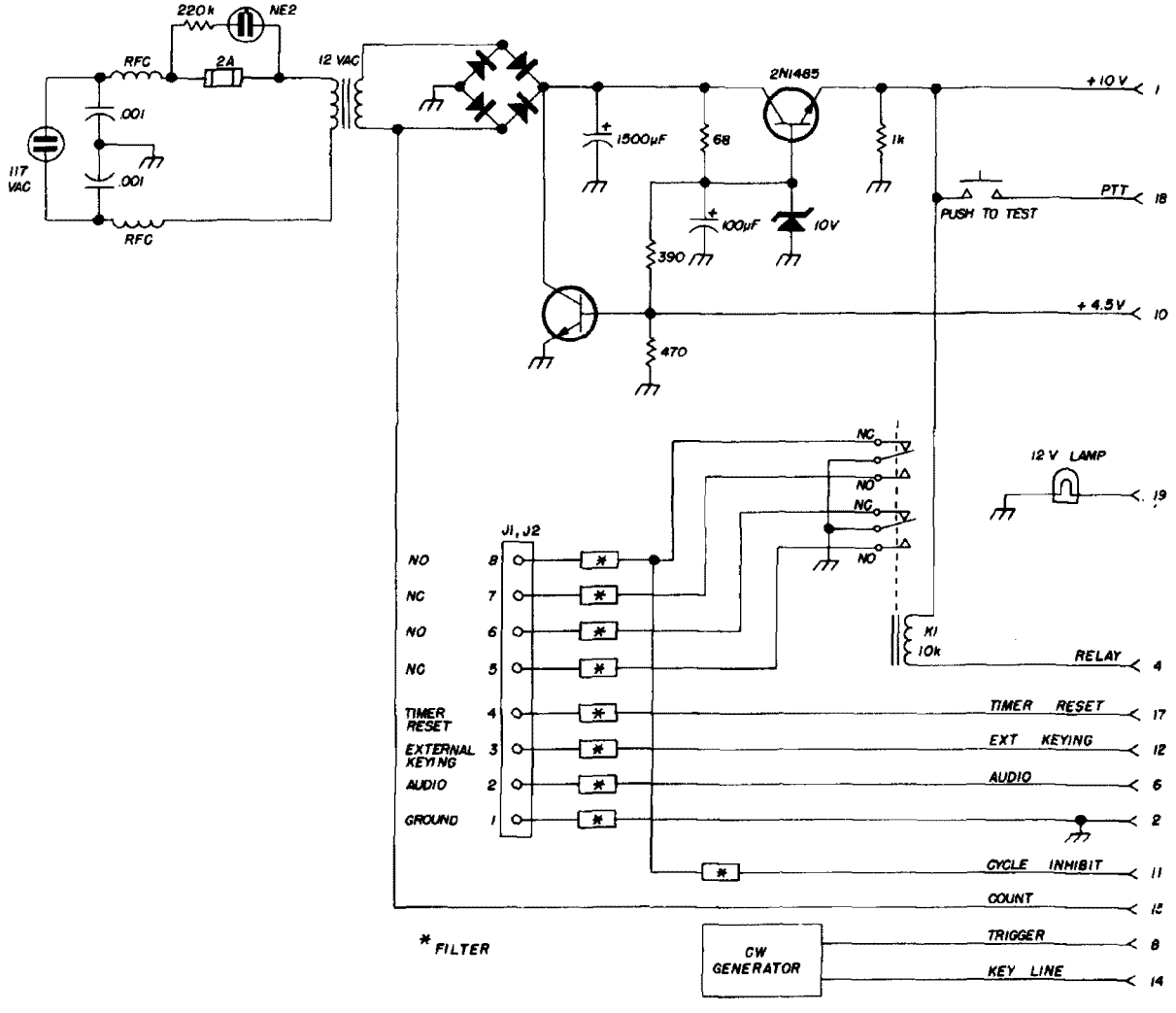
The big advantage is that by selecting the proper resistor values the code is chosen. The resistors could be mounted on a separate plug-in card and the code changed by changing cards.

logic plane

This card is the heart of the identifier. It provides all the control, audio, and keying functions. For clarity the schematic of this card has been divided into several sections. Each can best be understood by also referring to the block diagram **(fig. 3).**

The keying line output from the memory

fig. 1. Basic chassis contains the power supply, card plugs and interface circuits. Pins on the right mate with pins on the logic card.



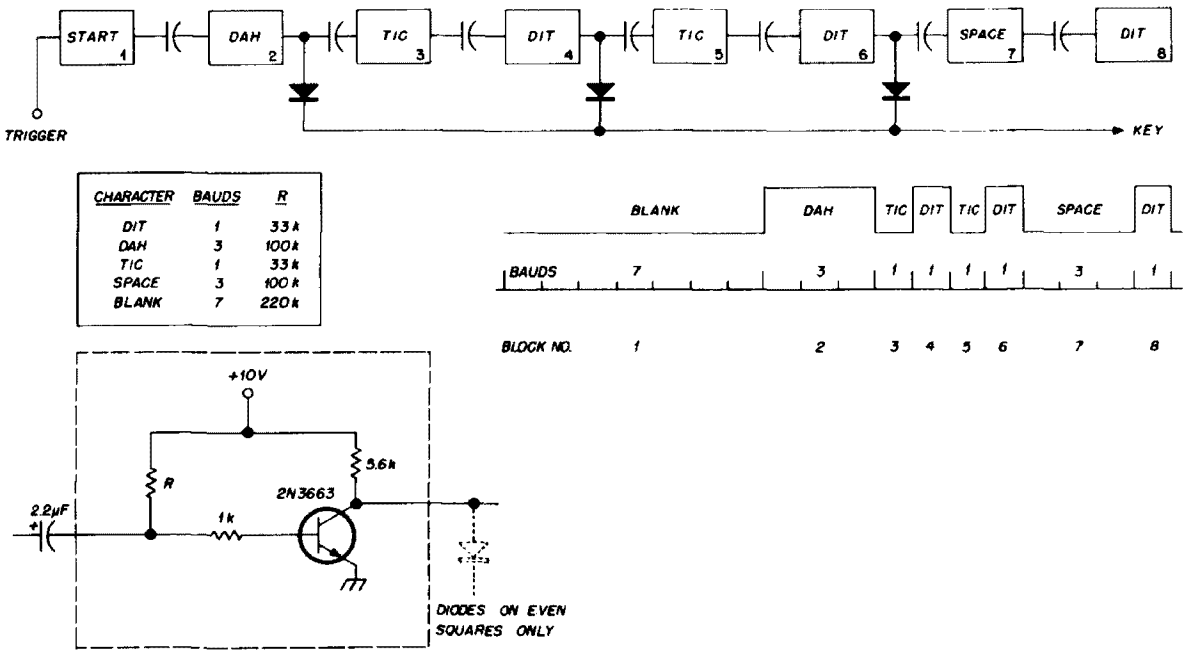


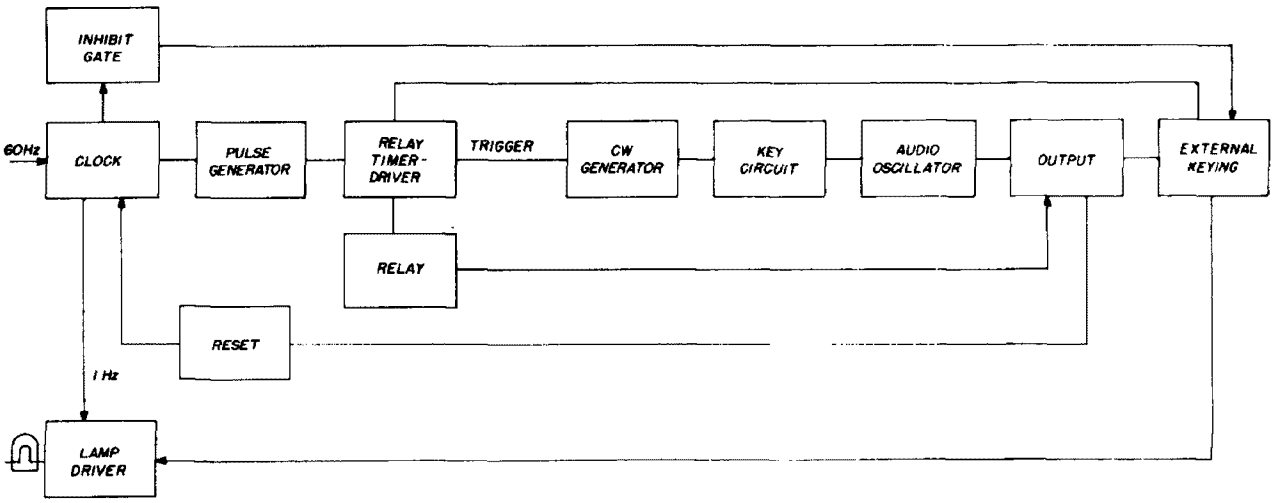
fig. 2. First eight stages of the CW generator or memory shown here generate the letters, "DE," as shown on the output wavetrain. Each of the boxes contains the simple transistor circuit shown. The length of the CW character generated by the circuit is determined by the size of the resistor R.

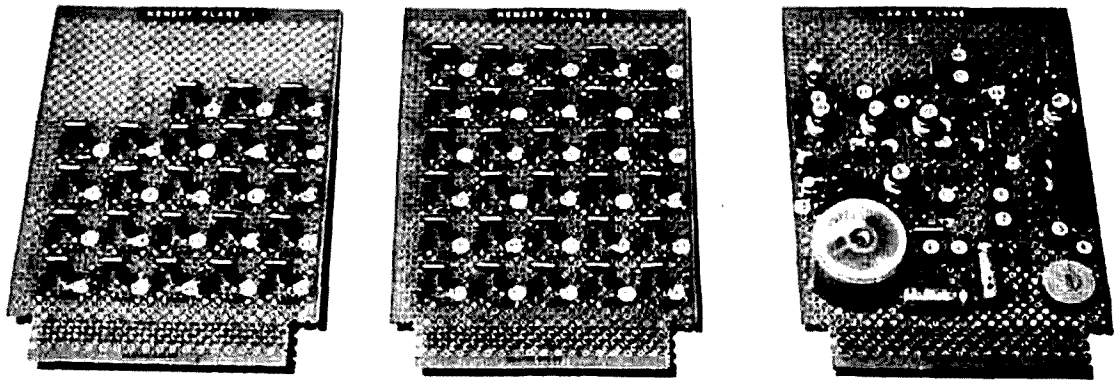
planes switches Q5 and Q6, which turn the audio on and off, **fig. 4.** Q1-Q4 make up a conventional oscillator. The 1k ohm potentiometer controls the output level. Sufficient audio power is available to drive almost any transmitter.

clock

The clock (**fig. 5**) establishes the ten-minute intervals necessary to fire the identifier. A 60-Hz, half-wave signal from one side of the power supply rectifiers is squared by Schmidt trigger Q23 and Q24 and applied to the clock input. The clock divides the line frequency by 36,000 to produce one pulse every ten minutes at point B. Q18 is the reset transistor, which allows the clock to be synchronized to ten-minute intervals. By listening to WWV while holding pin 4 of the output plug to ground and releasing it on the time mark, you can set the clock to ten-minute intervals on the hour. Using a

fig. 3. Logic plane is the heart of the unit: circuits are shown in figs. 4, 5 and 6.





Plug-in cards for the identifier: two memory cards on the left generate Morse characters; logic card on right provides the control, keying and audio functions.

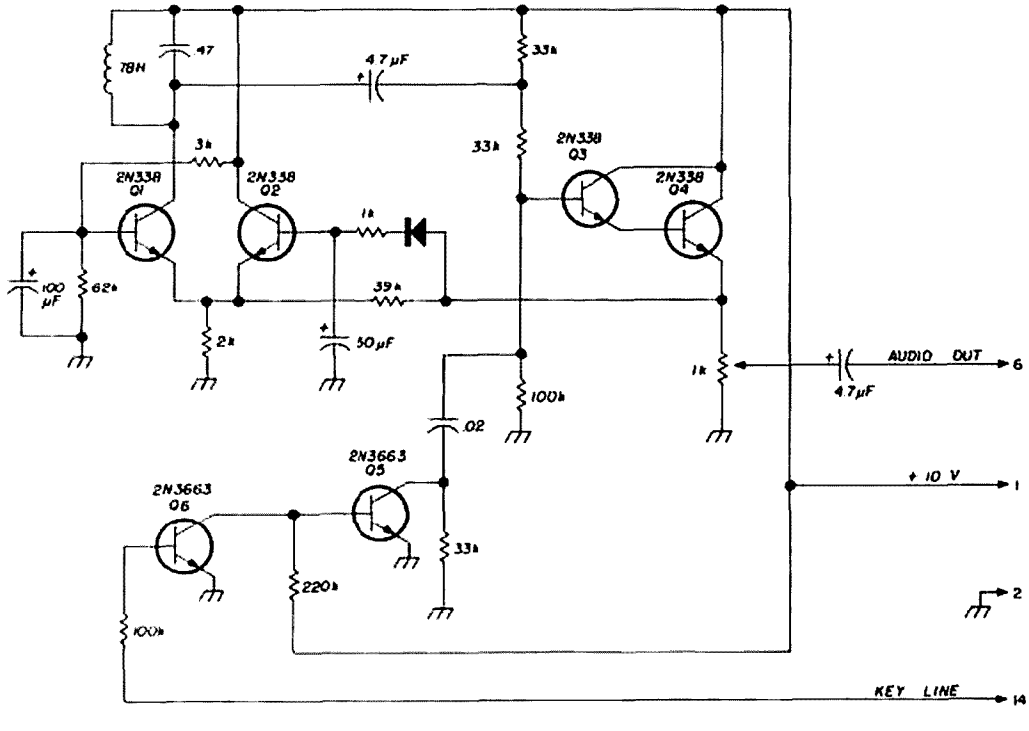
command from the command decoder allows the clock to be synchronized from a remote station.

access-control gate

This circuit is shown in **fig. 6.** The output pulse, at ten-minute intervals, drives Q16, which turns on the relay timer/driver composed of Q7-Q10. This closes the relay, which holds the transmitter on, and also provides the trigger pulse that starts the identification cycle. The one-second output from the clock is used to flash the lamp on the front panel at one-second intervals through Q11-Q13.

Point A is the takeoff for the one-second

fig. 4. Audio oscillator and keying circuit.

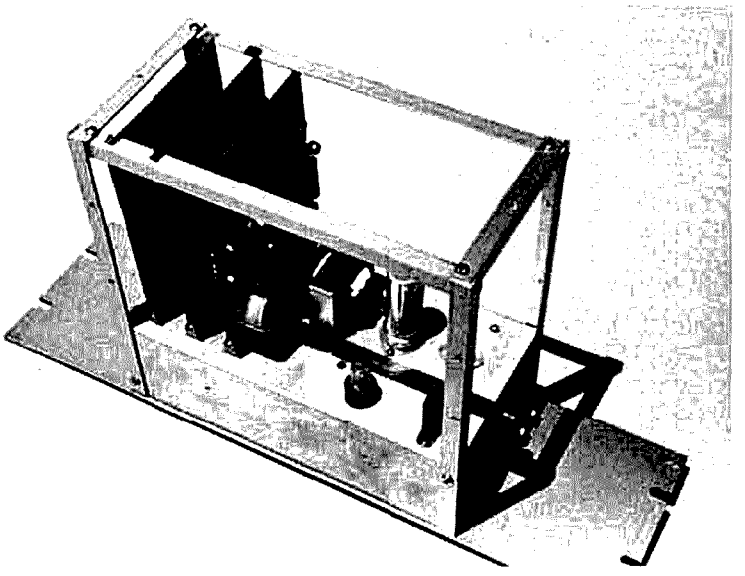


pulses. Q14 and Q15 are inhibit gates that prevent access to the trigger circuit from external sources such as the push-to-test button or the external keying input. There are only two times when it's necessary to prevent triggering the memory unit: first, when the identifier is cycling through it's sequence and, second, when the identifier is approaching a ten-minute interval on its clock.

The diode from pin 11 of the card plug clamps the inputs during cycling. The four diodes attached to Q14's base resistor sample the clock and clamp the inputs ten seconds prior to clock firing. At the same time, this clock-inhibit signal holds the panel lamp on constantly as an external signal to alert an observer.

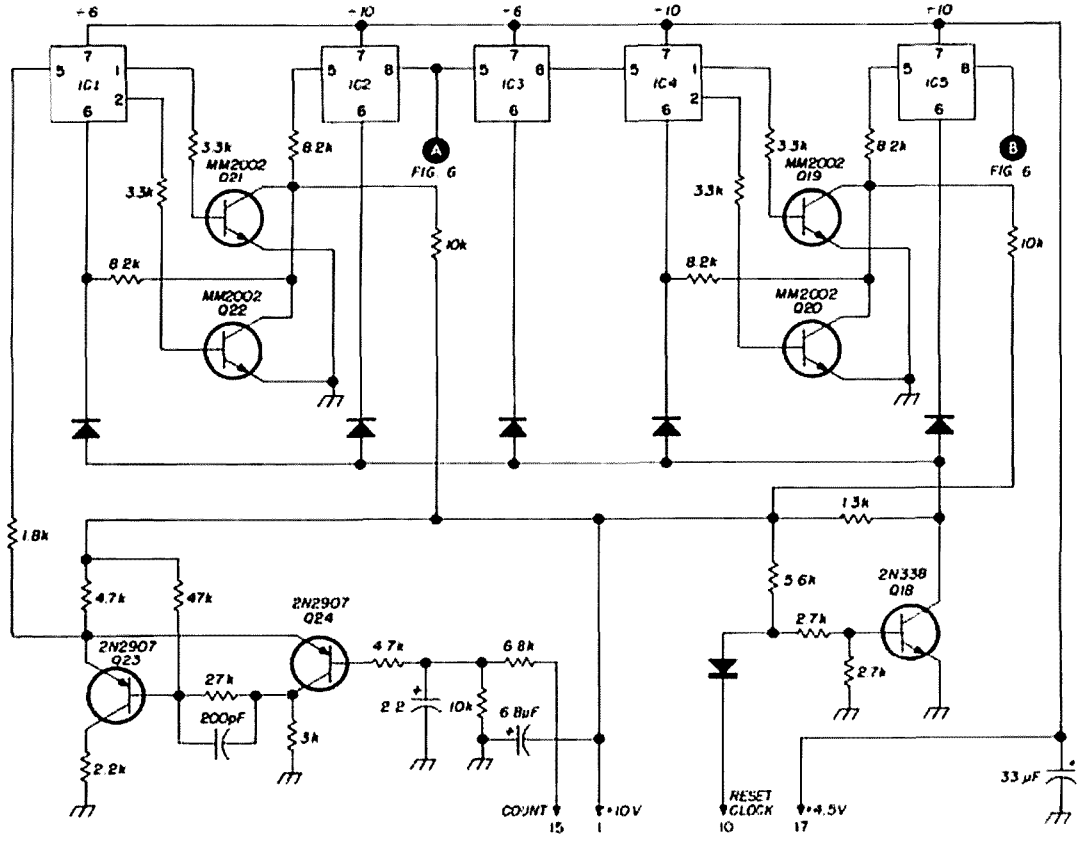
mechanical considerations

Repeater equipment requires a lot more attention to detail than most amateur equipment. It functions in a re-



Inside view of the CW identifier showing the three plug-in cards on the left and power supply to the right. The aluminum frame is made from do-it-yourself angle stock. BNC connectors are for audio and control functions.

fig. 5. Clock for the logic board derives its input from the 60-Hz line. Integrated circuits are Fairchild 995859's although other IC's may be used; divide ratios for each IC stage are shown.



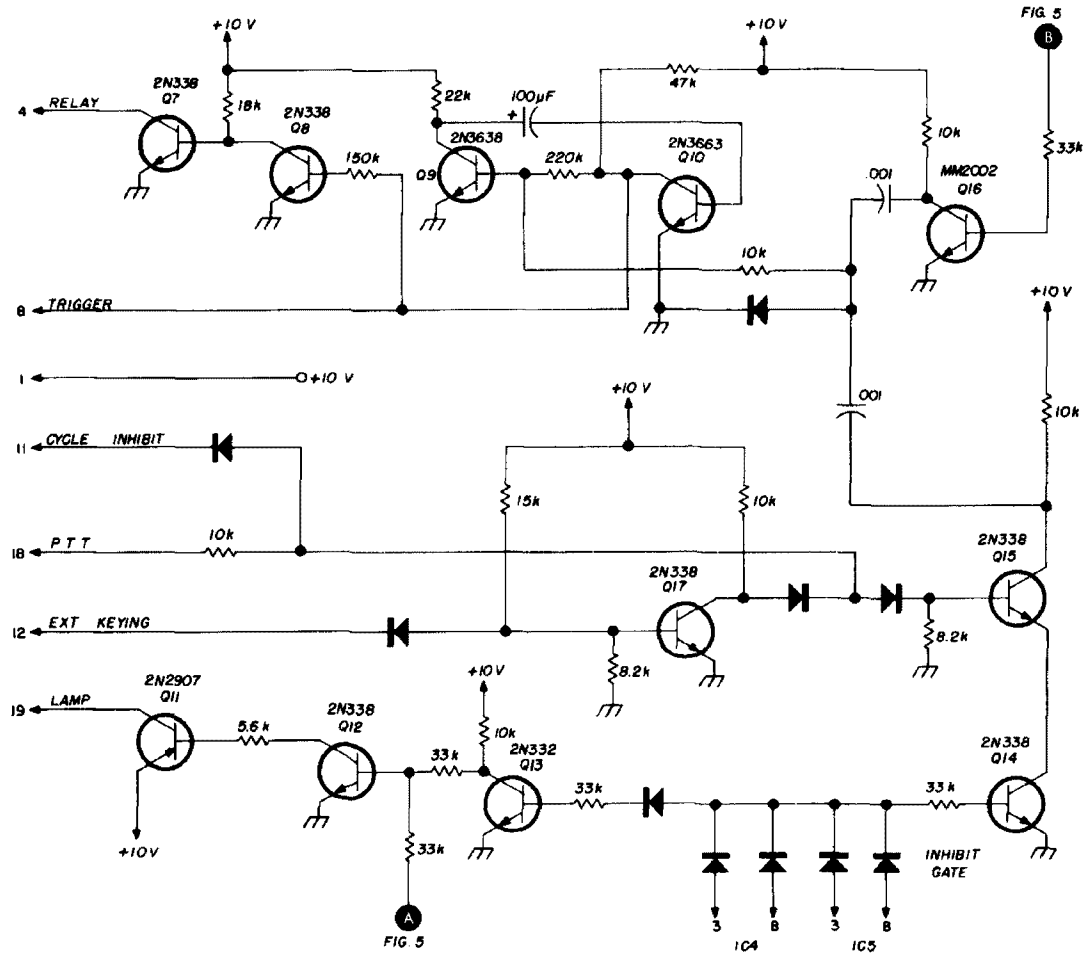


fig. 6. Access-control gate is used to start everything rolling at 10-minute intervals.

mote area where access is usually time-consuming and bothersome. The environment under which it operates is generally more hostile to solid-state equipment than a home or ham shack and often produces conditions totally unanticipated. For these reasons you should use care in constructing the identifier, and considerable attention should be paid to detail.

Numerous filter circuits are used to remove transients and rf, which would otherwise cause erratic triggering and operation. The photographs show the extent to which shielding was employed as a precaution against the intense rf fields from the repeater's transmitter. All input-output lines are filtered, including the 110-volt power line.

Extreme temperature variations are another problem more critical with transistorized equipment. Although this unit was heated in an oven and cooled in a refrigerator to test its range of temperature tolerance, I recommend it be installed with adequate ventilation and with no heat-producing elements directly under it. The best method is to give it a rigorous test before installing it at your site.

Even after trying everything I could think of to make my identifier act up, something totally unanticipated forced me to go up to the site and get it for modification. All-in-all, however, the problems were minor and quite easy to correct. The unit functions quite well now and should be in service for a long time to come.

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ham radio

to clip or not to clip?

That's the question– and here are some provocative answers

E. H. Conklin, K6KA, Box 1, La Canada, California 91011

Some of the amateur phone signals indicate that speech compressors and clippers aren't necessarily the answer to good communication. This article attempts to demonstrate that, under certain controlled conditions, speech compression and clipping will increase the weaker voice components. It further shows that when these devices are not used properly, a penalty is incurred in the form of noise. Noise is defined as an extraneous by-product of radio transmitters that causes interference to other amateurs. Also, the article shows that if the peak power of ssb transmitters is limited with speech clipping, the amateur regulations can be violated by exceeding the legal average input power limit.

speech compression

Years ago, an aviation engineer in military service said that at least four aviators jumped out of their aircraft as a result of misunderstanding something said to them over a radio circuit. While this may not happen to amateur operators, the desire is great for intelligibility of signals at distant points.

Bell System engineers state that the dynamic range for voice communication is 20 to 30 dB. A considerable amount of the received intelligence depends upon the weaker parts of the voice wave. This range above the inherent circuit noise normally is necessary.

Several decades ago the **compandor** was developed to reduce this limitation. The compandor compresses the audiody-

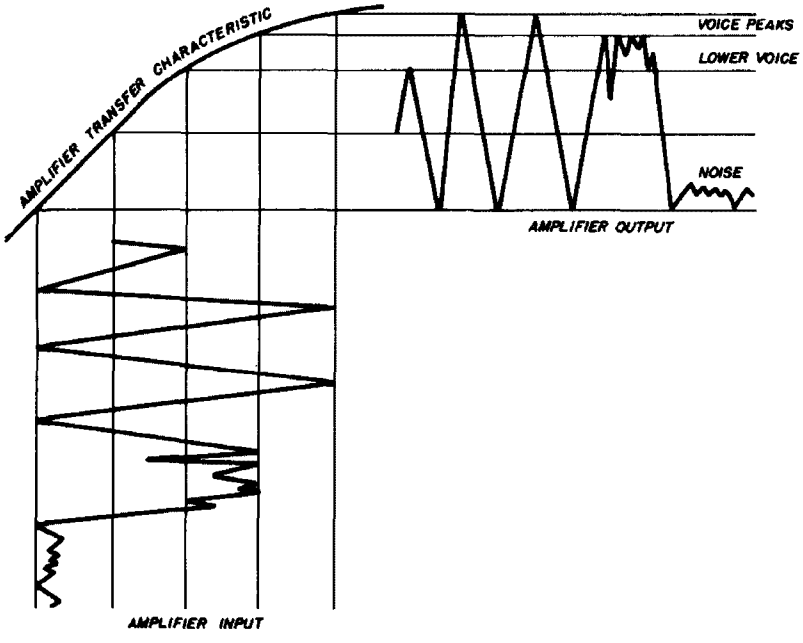
namic range before much noise is present and expands it again at the receiving end after transmission over a long circuit subject to noise. The combined action thus corrects most of the distortion created by the compression of the amplitude range at the sending end.

Compression of amplitude range might be pictured as in **fig. 1.** Here, an amplifier that is nonlinear above a point about equal to the lower amplitude components of the voice range will predistort and compress the range of the peaks above this level. The lower-amplitude components remain at the same level above the noise. This performance is not greatly different from filtered audio clipping, which attempts to do the same thing by clipping the peaks and then rounding them by filtering the high frequencies generated in the process. Either of these operates on a cycle-to-cycle basis.[1]

Another type of compression is much more common. It operates on the integrated audio waveform principle, having an effect similar to automatic gain control. Its advantage comes from keeping the gain at some predetermined high level, even when the mouth is turned from the microphone. Its benefits are small.[1, 2, 3] Some form of indicator can be used, instead, to maintain a standard voice level.

One disadvantage of the integrated-waveform compressor is the tendency for hum, wind (breath) noises, blower, background noise and even tube noise to increase during pauses in speech.[4] It causes some operators to hurry their speech, which is then difficult to interpret. This form of compression results in a sharply rising noise level after speech ends and before the VOX or push-to-talk circuit cuts off the transmitter. It often sounds like interference.

fig. 1. Compander performance, showing cycle compression of peaks.



single sideband

A peculiarity of ssb is that occasional peaks are generated. [1, 3] The accepted theory is that, if these peaks are reduced in amplitude by rf clipping and filtering, they permit the audio level of the weaker voice components to be increased without greatly reducing the intelligibility, because the peaks contribute little to the intelligibility.[5] At a distant receiver where the noise level may be close to that of the lower-amplitude voice components, intelligibility may have suffered considerably. An increase in audio may improve the reception more than distortion causes it to deteriorate.

This might be depicted as in **fig. 2.** The weaker voice components are in the noise at the distant receiver whereas with the same peak level, clipping may allow the weaker components to increase at the transmitter. Then they will be above the noise at the receiver.

local interference

This rationale overlooks the interference that may be generated locally, which may cover the entire band. Some amateur signals indicate that many transmitters were modulated 100 percent with noise, after which the voice simply punched holes in this noise. A later inspection of a maintenance activity disclosed that the original Collins design was being produced by other manufacturers. The Collins technical note was not being distributed to all who should have it to prevent the accidental setting of the audio gain at a too-high level.

A later study of ground/air communications by the Naval Research Laboratory indicated aircraft radio sets were only 25 percent intelligible on the average, and de-

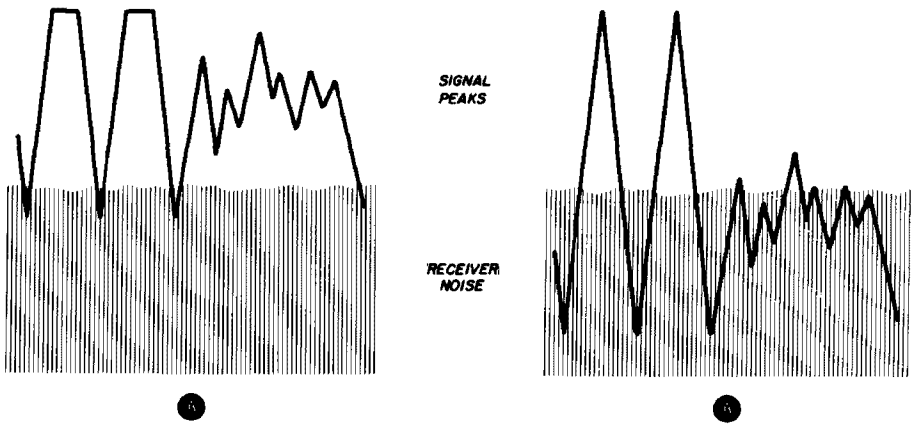


fig. 2. Signal received at a distant receiver when (A) clipping allows transmitter power on lower voice components to be increased, assuming that this brings the weaker parts of the voice above the noise. In B the receiver noise is above the lower voice components.

should not use either compressors or clippers, at least as they're used at present.

A local amateur told me, "If you don't use a clipper, you're not up to the state of the art." This latter term is used in section 97.73 of the amateur regulations: "The frequency of the emitted carrier wave shall be as constant as the state of the art permits." The implication is that not only shall the carrier wave frequency be as constant as the state of the art permits, but distortion products shall be eliminated in accordance with "... good engineering and good amateur practice" (from section 97.77).

intelligibility

During the invasion of Okinawa, I was impressed with the intelligibility of aircraft radio signals. A scope showed they depended upon redundancy (familiarity with what probably would be said) for communication. This was largely due to the improper setting of the gain control, which produced excessive noise in the output. The report also showed the ground stations were only 50 percent intelligible during carefully controlled tests. To keep modulation at high levels, they were using compression plus some clipping. When this was turned off, the problem disappeared.

an example

When an amateur ssb signal swings an s-meter up, and the meter doesn't vary more than 20 or 30 dB between words, noise is being transmitted, which causes interference. This is not necessarily radiated beyond the passband of the trans-

mitter filter, but a noisy signal is being radiated that can interfere with the whole band locally, and with communication between other stations on approximately the same frequency.

Let's say X is talking to Y on low power, some distance away from Z, and not interfering with Z. But Z's noise may well prevent Y from hearing X. X's signal at Y might look like **fig. 2a** when Z is off, and like **fig. 2b** when Z is transmitting. This is not desirable and should be avoided as unnecessary to Z's communication.

Oscilloscope observation of some of the compressed or clipped signals shows that the voice is punching holes in the noise just as the Okinawa aircraft sets did nearly 24 years ago. Others are clean, allowing *weaker signals to be heard* through a louder one. Sometimes, a rare station stops transmitting, after which so many call him that the band sounds like escaping steam. This has been attributed to the large number of stations calling, but may well be the result of a single transmitter that's producing noise. A New Zealander once tried to describe this condition diplomatically to an East-Coast station one evening when a breaker said, "He sounds like a one-man pile-up!" At that, the ZL nearly blew a fuse!

peak power

Some operators are impressed with the higher average plate current that results from the use of a compressor or clipper, thinking that the higher current results from the voice. This isn't necessarily so. Also, the legal limit will be exceeded unless the power is reduced when this occurs. Tests with and without a clipper are difficult to control on a truly comparative basis, so a report of a better signal with the clipper may be only the result of an illegal power input.

A single sideband transmitter can put out considerable power without even moving the plate-current meter above the static level, because the meter time constant won't respond to transients (noise pulses and shot noise, for example). If a compressor or clipper causes the plate current to rise without voice modulation, clearly the compressor or clipper is not operating properly and is modulating the transmitter with noise.

A single-sideband transmitter can put out considerable power even without moving the plate current meter above the static level. The meter may move as little as 6 dB from the static level to the point coinciding with 1kW output. Therefore, between words, the driving power should be far less than that which would just start to move the meter. If a compressor or clipper causes the plate current to rise without voice modulation, clearly the compressor or clipper is not operating properly and is modulating the transmitter **heavily** with noise.

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4. Sydney H. Moate, "An Audio Peak Limiter for Voice Transmission," *QST*, Vol. 49, No. 8, August, 1965.
5. "RSGB Handbook," Third Edition, p. 278.

ham radio



"Why, hello Marilyn. Looks like you put on a little weight. How are the kids? Did my five-band DXCC plaque arrive yet?"

a survey of high-frequency amateur antennas

With special emphasis on a new family of antennas that offers superior performance and wideband operation

George A. H. Bonadio, W2WLR, 373 East Street, Watertown, New York 13601

Every amateur has different antenna needs, different facilities and a different pocketbook, so no one antenna will satisfy everyone. Indeed, it is the wise operator who considers many types of antennas before he decides what he is going to use. The chart in **table 1** was prepared as a guide to help you visualize some of the important differences between different types of antennas found on the amateur bands. Hopefully, it will help you make that all-important decision regarding your next sky wire.

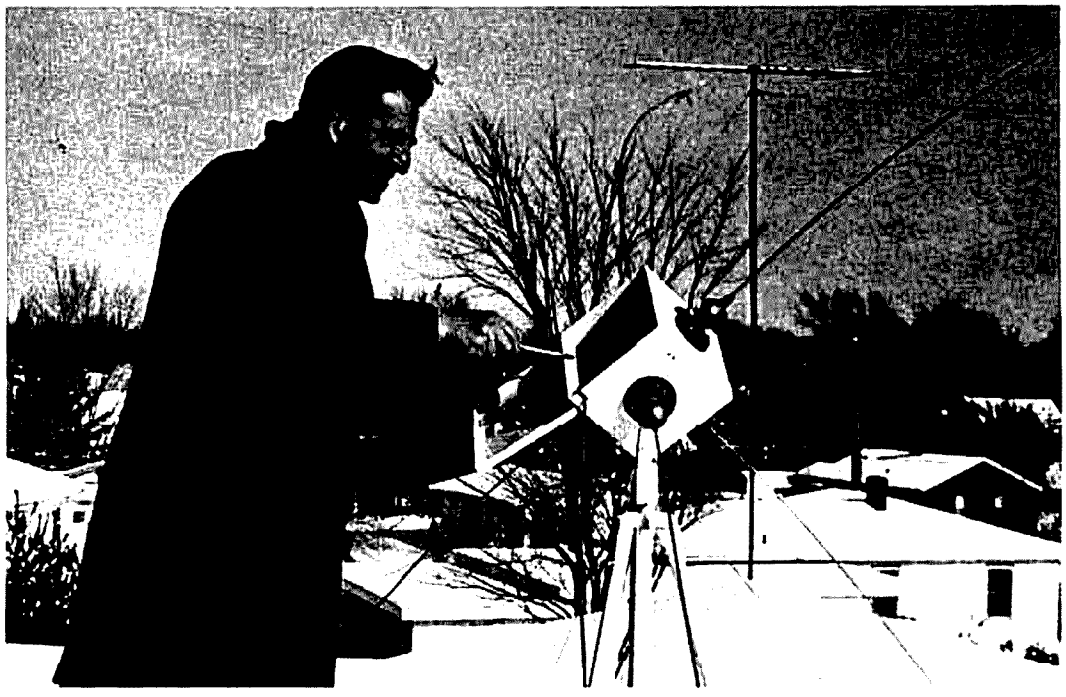
bonadio antennas

You should be familiar with the first fourteen antennas on the chart, but the last four are new inventions* with some very interesting and exciting properties. Most antennas radiate basically along one vector. Exceptions include the Marconi, the quad and the rhombic; these have low- to medium-grade two-vector radiation. The Bonadio square-diagonal antenna radiates a high-grade two-vector wave. It does this by equally dividing the radiated energy into two identical members, 90° apart. The Bonadio box-diagonal antennas carry the idea one step further and divide the energy into four equal members 90° apart.

The Bonadio space-dimension antenna (**fig. 4**) completes the separation into idealized waves 90° apart. A logical presumption at this point is that such thin-

* The Bonadio square-diagonal antenna is protected by U. S. Patent 3,274,606. Patents are pending on the Bonadio box-diagonal and cube-diagonal antennas; Mr. Bonadio has applied for a patent on the space-dimension antenna. This protects the inventor against commercial infringement but does not prevent an amateur from building one for his own use. **Editor**

ning of the radiation patterns is wasteful, but **results** do not confirm this. On band openings and closings, in fact, the extra wave vectors of the transmitted energy appear to be associated with the performance I've experienced with them. The following reports are typical of the ones I receive when using Bonadio antennas, and suggest a number of unexplained factors. Just about all the aspects of my



W2WLR and the latest model of the space-dimension antenna.

signal being the first in **and** the last out. In addition, numerous spontaneous reports of less fading are logged. Comparisons with Yagis, quads and rhombics show more advantages at the more distant locations and the advantages seem to accumulate with each additional ionospheric hop.

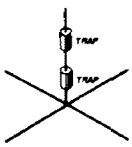
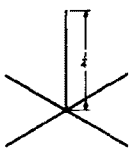
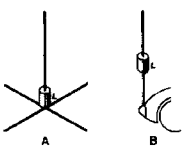
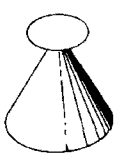
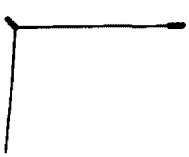
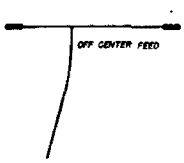
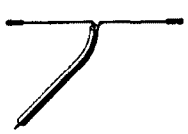
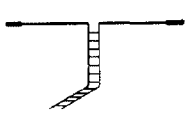
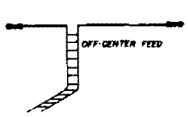
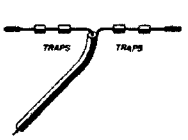
The performance of these antennas may be related to the three-dimensional nature of the ionosphere. The ionosphere **appears** to reflect two-vector waves with less fading and absorption than one-vector waves; furthermore, it seems to reflect three-vector waves with less fading and absorption than waves arriving along two vectors.

results

However, the main purpose of this article is not to try and explain how these antennas work, but rather to present the

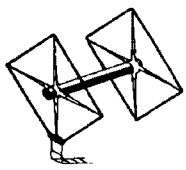
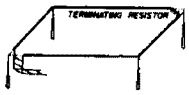
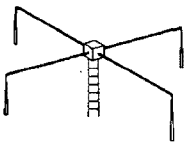
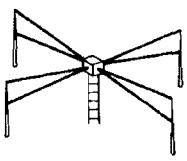
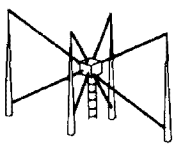
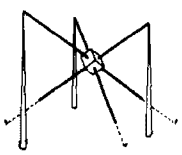
As you read through this article you will note a number of unusual and even unorthodox uses of the words "vector" and "wave." This is because of the apparent nature of these antennas. The author originally referred to the radiation patterns as two- and three-dimensional, but all antenna patterns have vertical height and horizontal width and depth, so I pointed out that a three-dimensional radiation pattern is not unique. The word "direction" could have been used, but many antennas are multi-directional or have several lobes, so this is not unique nor descriptive of what the author feels is happening. Finally, after much discussion, the word "vector" was chosen because it means, "along a single line." Please bear in mind that the author is not certain how these antennas work, although he has some ideas. Performance is what is important, and these antennas have that; we'll leave it to the antenna engineers and Maxwellian mathematicians to explain why the performance is there. Editor

table 1. Detailed comparison of 18 different high-frequency amateur antennas.

	type	ground wave	short skip	long skip	comparative cost	support points	ground system	short term fading	use for cb, tv fm, uhf etc.	feeder losses
	vertical trap	excellent	poor	good	medium	1	very important	very poor	very poor	low
	vertical ¼-wave	excellent	poor	good	low	1	very important	very poor	very poor	low
	vertical, shortened and loaded	very good	very poor	fair	high	1	very important	very poor	very poor	low
	discone	excellent	poor	good	medium	1	part of system	very poor	very poor	low
	Marconi, long wire	good	good	fair	very low	2	modest	good	good	part of system
	Windom single wire	fair	very good	fair	low	2	modest	poor	fair	medium
	doublet, coax fed	good	very good	fair	low	2	none if high up	poor	very poor	medium
	doublet, open wire fed	poor	very good	fair	low	2	none if high up	poor	fair	very low
	doublet, Windom feed	poor	very good	fair	low	2	none if high up	poor	fair	low
	trap doublet	poor	very good	fair	medium	2	none if high up	very poor	poor	low
	inverted-V doublet	fair	very good	fair	low	1	none if high up	good	fair	low

optimum results on	tuning requirements	time to change pattern	pattern polarization	ground space not including guys	anti-QRM notch	anti-line noise ability	type
several narrow bands	fixed	—	vertical	ground radials	no	vertical trap	none
one band	fixed	—	vertical	ground radials	no	none	vertical ¼-wave
one narrow band	special loading	—	vertical	ground radials	no	none	vertical, shortened and loaded
10:1 frequency range	none	—	vertical	circle	no	none	discone
any frequency with tuner	medium	hours	mixed	straight line	no	very poor	Marconi, long wire
one band	simple	hours	horizontal	straight line	no	poor	Windom single wire
one narrow band	none	hours	horizontal	straight line	no	fair	doublet, coax fed
any tuned band	modest	hours	horizontal	straight line	no	fair	doublet, open wire fed
any tuned band	modest	hours	horizontal	straight line	no	fair	doublet, Windom feed
several narrow bands	fixed	hours	horizontal	straight line	no	fair	trap doublet
narrow band	*fixed*	*hours*	*mixed*	*straight* line	*no*	*poor*	inverted-V doublet

tabla 1. Detailed comparison of 18 different high-frequency amateur antennae.

	type	ground wave	short skip	long skip	comparative cost	support points	ground system	short term fading	use for cb, tv fm, uhf etc.	feeder losses
	Yagi, 3 element	good	very poor	excellent	high	1	none	poor	very poor	medium
	quad, 2 element	very good	poor	excellent	high	1	none	very good	poor	medium
	terminated rhombic	fair	poor	superior	high	4	none	good	good	very low
	Bonadio square-diagonal	poor	very good	good	medium	2-4	none if high up	very good	excellent	very low
	Bonadio box-diagonal	fair	excellent	very good	high	4	none if high up	excellent	excellent	very low
	Bonadio cube-diagonal	excellent	excellent	excellent	very high	4	none if high up	excellent	excellent	very low
	Bonadio space-dimension	excellent	excellent	excellent	very high	3-6	none if high up	superior	excellent	very low

station and location are typical—140 watts (a-m), located in a built-up residential area, five-degree hill toward Africa—so the exceptional reports I get aren't from that department. The antenna in use when the following reports were received was a Bonadio box-diagonal antenna, elevated 5 to 9 meters (16 to 30 feet) above flat land.

K5MRU, 16 September 1967, "There must have been 15 stations in there calling me."

ON4XK, 16 September 1967, "There were many stations on your frequency."

ZE1BP, 26 September 1967, "You're on a par with **any** stateside signal in the last hour."

ZS6EB, 27 September 1967, "You're the **only** station we can copy," and four hours later, "Conditions no better than earlier; you're still 4 and 5." And on 5 October 1967, "First W2 heard this afternoon. There are no Americans; the band is no good at all. You're 4 and 6."

G3RXW, 16 October 1967, "About the only W on the band;" 40 minutes later, "I cannot hear any other American station."

optimum results on	tuning requirements	time to change pattern	pattern polarization	ground space not including guys	anti-QRM notch	anti-line noise ability	type
single band	fixed	seconds	vertical or horizontal	depends on support	excellent	excellent	3-element Yagi
single band	fixed	seconds	mixed	depends on support	excellent	excellent	2-element quad
3:1 frequency range	fixed		mixed	diamond acreage	no	fair	terminated rhombic
all above half-wave resonance	special tuner	immediate	vertical or horizontal	large	poor	good	Bonadio square-diagonal
all above half-wave resonance	special tuner	immediate	mixed	large	poor	good	Bonadio box-diagonal
all above half-wave resonance	special tuner	immediate	vertical or horizontal	large	poor	good	Bonadio cube-diagonal
all above half-wave resonance	special tuner	immediate	vertical or horizontal	large	good	very good	Bonadio space-dimension

F3EA, 17 October 1967, "Your signal is one of the very best I can hear on the band; very little QSB."

G3OSI, 20 October 1967, "When I heard you calling 'CQ DX,' you were the strongest on the band, but there is an ssb stronger."

ZS1JH, 24 October 1967, while in a four-continent roundtable with G3HCU and the late ZL2UD, "This evening there is fading on all signals, but **not** on your signal, George."

From G3HCU, during the same roundtable, "George has a signal we classify as outstanding here in England."

Note that some stations are running up to 8 dB more power (kilowatt input), plus 10 dB or so for antenna elevations and up to 8-dB beam-antenna gain; this is more than 26 dB over W2WLR*—a 400:1 power ratio or more than four S-units. I

* And this does not include the theoretical 9-dB gain offered by single sideband.

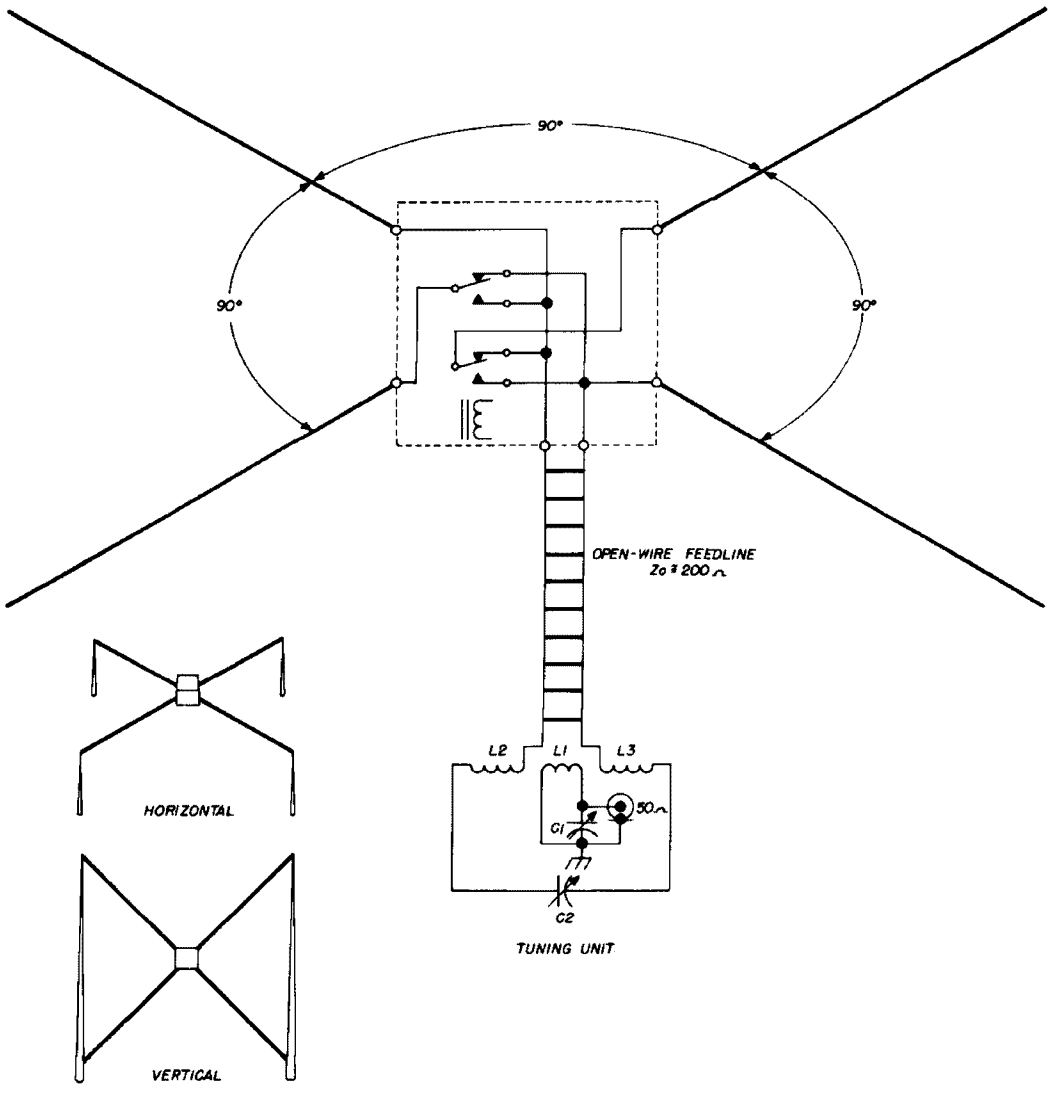


fig. 1. The square-diagonal antenna. The switching relay in the center is a modified dpdt unit; the special tuning unit is very important to proper antenna operation.

don't know what happened to the signal, but it is not happening at ground level; it must be happening in the ionosphere.

While splitting up antenna power three ways results in nearly 5-dB "loss" in the desired direction, it appears to mitigate fading depth by as much as 25 dB. Also, selective fading garble seems to be overcome by these antennas.

square-diagonal antenna

As the name implies, the square diagonal antenna shown in **fig. 1** is layed out along the diagonals of an imaginary square. It can be mounted in the horizontal plane on four supports or in the vertical plane on two poles. Both of these installations are shown in **fig. 1**. Each of the legs of the cross are exactly the same length and are made of number 8 aluminum ground wire or number 12 copper wire.

The relay in the center is a dpdt relay that has been modified for minimum lead length. The special antenna tuner shown in **fig. 1** is extremely important to proper operation and must not be left out. The reactance of L1 and C1 should be equal to one-third the impedance of the coaxial feedline if they are in parallel (three times the impedance of the coax if in series).

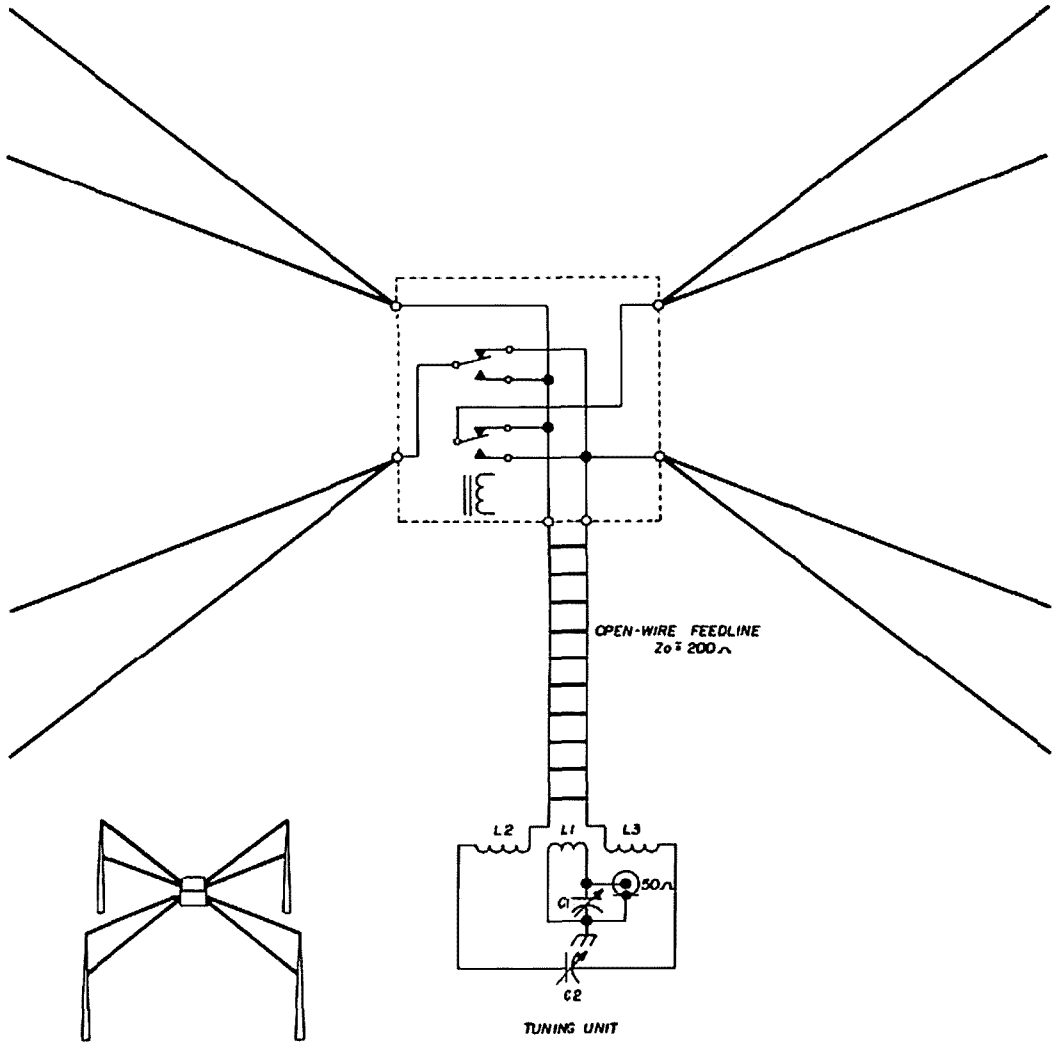
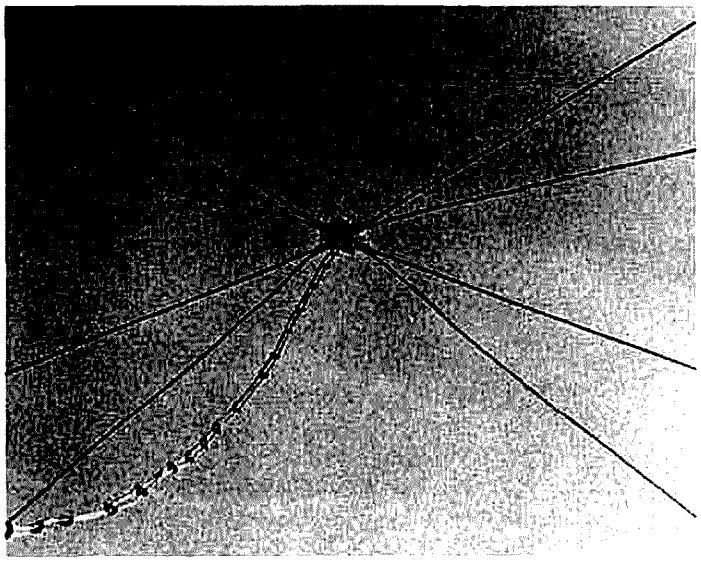


fig. 2. The box-diagonal antenna. The switching relay and tuning unit are the same as fig. 2.



Box-diagonal antenna.

The L1-C1 tuned circuit is grid dipped to the middle of the band before the coax cable is connected and **not** retuned. The reactance of L2 and L3 is three to five times the value of the impedance presented at the tuner by the open-wire feedline. If L1 is closely coupled to L2 and L3, an swr of 1:1 is an indication that the proper LC ratios have been used.

box-diagonal antenna

The box-diagonal antenna shown in **fig. 2** is a natural development of the square-diagonal antenna. The special switching relay in the center and the tuning unit are the same as those used with the square-diagonal. The length of each

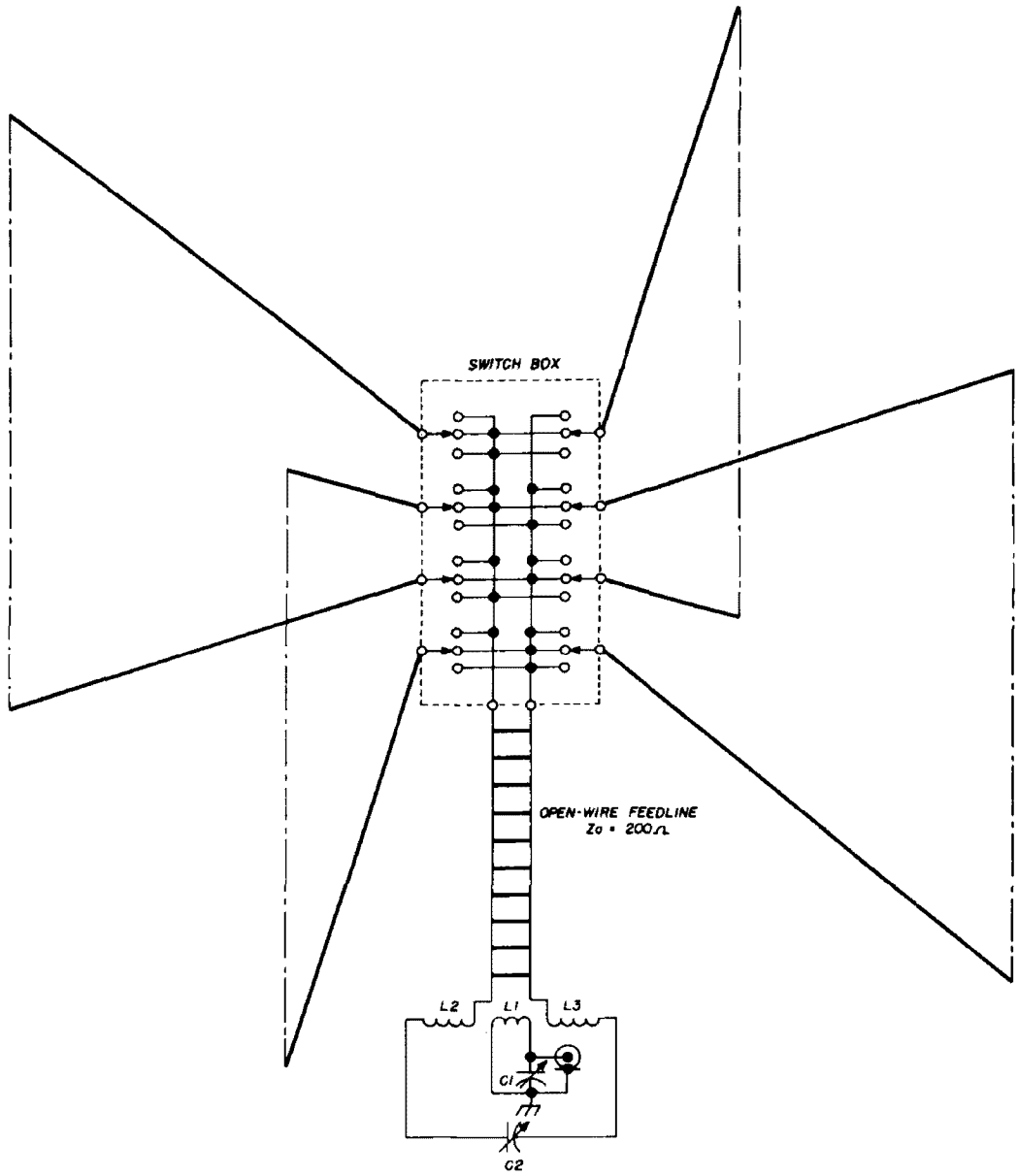


fig. 3. The cube-diagonal antenna. The element ends are located at the corners of an imaginary cube.

of the eight elements is exactly the same as before, and the vertical spread between wires on all four masts is identical. As you can see from **table 1**, this slight refinement in antenna construction results in improved signal reports—both close in and at long distance—as well as improved short-term fading performance.

cube-diagonal antenna

The cube-diagonal antenna shown in **fig. 3** is the result of efforts to further refine this family of antennas. The eight elements are exactly the same length and are terminated in insulators located at the corners of an imaginary cube. This antenna is somewhat similar to the box-diagonal, but note that the switching relay in the center is much more complex than the relay used in previous designs.

If you trace out the switching positions of the relay, you will find that in the center or neutral position the four upper elements are fed against the four lower elements. In the lower position of the relay the front four elements are fed against the rear four; and in the upper position, the left elements are fed against the right.

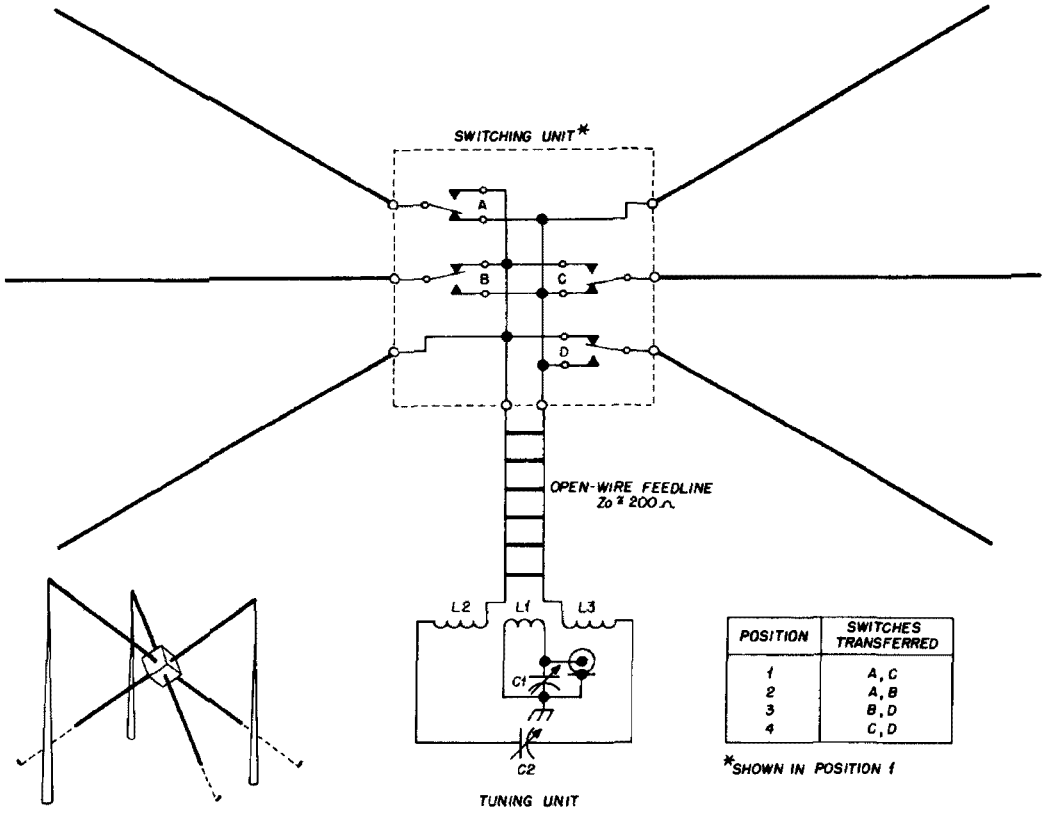


fig. 4. The space-dimension antenna; high performance with three support masts.

There was no commercial relay available to do this job, so a special unit had to be built.

space-dimension antenna

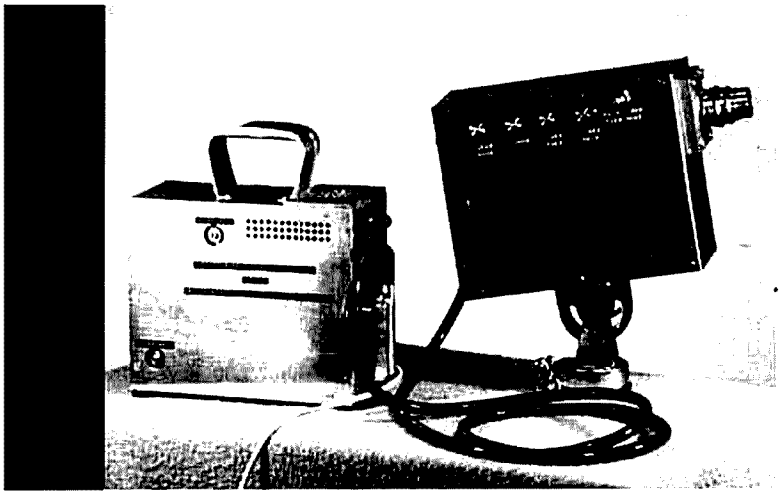
The space-dimension antenna shown in **fig. 4** is the latest effort to improve performance and simplify construction. Only three vertical masts are required for this antenna, but it out-performs all of the other Bonadio antennas. All six elements of the space-dimension antenna are exactly the same length and constructed of number 8 aluminum or number 12 copper wire; all angles are either 90 or 180 degrees.

Four switching positions are required at the center of the antenna to provide one vertical and three horizontal polarization patterns while maintaining correct tuning and loading; if you trace out the switching logic, you'll see that three adjacent elements are fed against the other three. It would be possible to construct a single remote-control switch to handle the complex switching chores, but the same results can be obtained with four spdt relays energized as shown in the table in **fig. 4.** The optimum elevation of the switch above ground is 20 meters (66 feet).

summary

Although these antennas have been in a continuous state of evolution since 1953, a lot of work remains to be done—experiments at different locations, performance tests at different frequencies and under different operating conditions, vhf and uhf tests to determine performance over line-of-sight paths—this is just a sample of the kinds of data that are needed. Construction of the antennas themselves is not difficult, although the complex switches and high-power-factor antenna tuners must be used for proper performance. Look for complete construction details in **ham radio** soon.

ham radio



new solid-state camera and monitor for slow-scan television

These new solid-state designs from Canada use sampling techniques to provide slow-scan pictures with fast-scan cameras

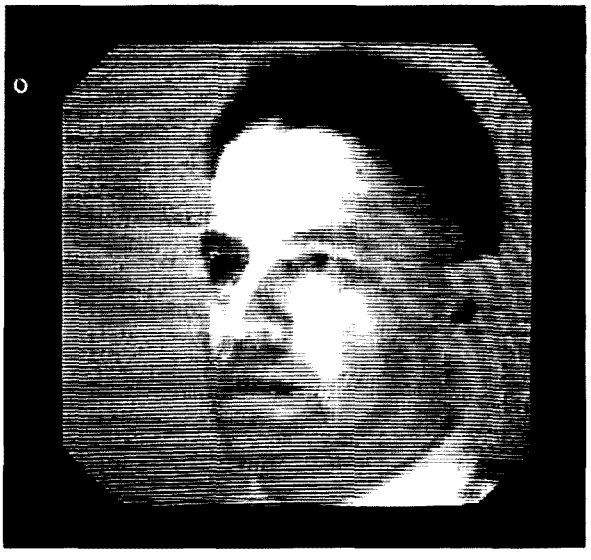
D. J. Watson and S. M. K. Horne, VE3EGO, Northern Electric Co., Ltd. Research & Developments Labs, P.O. Box 3511, Station C, Ottawa, Canada

The possibility of transmitting pictures has probably appealed to radio experimenters since the early days of radio. Hugo Gernsbeck featured the "teleautograph" on the cover of one of his early publications before 1910, and a cartoon in the May, 1921 issue of **QST** portrayed father tweaking the controls of his "radiotelescopograph." When practical (and realistically priced) hardware became available in the mid-1930's, radio amateurs were ready and the pages of **QST** were filled with tv technical and construction articles—so many in fact that some readers complained.

Many over-the-air friendships have been established between amateurs who have no idea what the other looks like, so the underlying interest in television should come as no great surprise. Unfortunately, normal fast-scan television requires wide bandwidths that are not compatible with our relatively narrow high-frequency allocations. However, slow-scan television—perhaps more properly called narrow-band image transmission—makes it possible.

In 1957, Cop MacDonald, now WAØNLQ, became intrigued with the possibility of using a slow-scan technique for image transmission by radio. This method involves the use of television-type pickup and reproduction devices with slow scanning rates to produce narrow bandwidth video signals. In September, 1957, he started the design and construction of a low-cost slow-scan system for use with amateur phone equipment.[1]

The system required no greater bandwidth than normal voice communications and the signal could be transmitted and received with any equipment suitable for radiotelephone work. Ideally, the "camera" could be plugged into the microphone jack of the ssb transmitter and a "monitor" plugged into the headphone jack of the receiver to convert a modern amateur radio rig into a two-way picture transmission facility.



The authors via slow scan: Doug Watson. . .



and Syd Horne, VE3EGO

Along with WAØNLQ, Don Miller, W9NTP, J. A. Plowman, G5AST, Louis I. Hutton, K7YZZ, and others did much to pioneer sstv development by proposing standards, writing technical and construction articles and interesting other amateurs in slow-scan work. A number of on-the-air experiments were conducted on the bands below 30 MHz under special temporary FCC authorization; similar arrangements were worked out with the DOT in Canada. In 1961 the ARRL requested the FCC to permit narrow-band picture transmissions in the phone portions of the 10- and 15-meter bands; in 1967 the FCC issued a Notice of Proposed Rulemaking that would authorize narrow-band picture transmission in the lower frequency amateur bands; this was eventually adopted effective August 30, 1968. The slow-scan frequency allocations for the United States and Canada[4] are listed in **table 1.**

Syd Horne, VE3EGO, built the camera and monitor described by Copthorne MacDonald in QST.[5,6] After many months of work and planning, he exchanged pictures across the Atlantic for the first time with Art Backman, SMØBUO, in Stockholm.[7] But before this important milestone could be achieved, in addition to constructing the cameras and monitors, permission had to

This article is an adaptation of an article that originally appeared in the January, 1969 issue of electron—the Canadian electronics magazine edited by Ernie Welling, VE2YU—and is presented here through the co-operation of Tomar Publications, Ltd.

table 1. Slow-scan television frequency allocations below 30 MHz.

United States[1]		Canada[2]
present	22 November 1969	
3.800- 3.850 MHz	3.800- 3.900 MHz	3.725- 3.750 MHz
7.200- 7.225 MHz	7.200- 7.250 MHz	7.150- 7.175 MHz
14.200-14.235 MHz	14.200-14.275 MHz	14.175-14.350 MHz
21.250-21.300 MHz	21.250-21.350 MHz	21.100-21.450 MHz
28.500-29.700 MHz	28.500-29.700 MHz	28.100-29.700 MHz

1. Allocations limited to Advanced and Extra-class segments of the bands below 28 MHz.

2. Subject to revision on March 31, 1969.

be obtained from the licensing authorities in both countries to transmit sstv.

While operating with this equipment, Syd became aware of many of the operating restrictions that were imposed by existing equipment. He discussed the problem with Doug Watson and managed to get him interested enough in slow-scan tv to attempt a solid-state camera and monitor design incorporating features which would facilitate operation. The camera prototype was completed in June, 1968, and the monitor in October, just in time for a sstv demonstration at the first annual Radio Society of Ontario convention. With these new designs, the camera **samples** the output from a fast-scan picture source, producing pictures at the slow-scan rate.

the sampling camera

The camera uses a regular vidicon, such as the 7038, instead of the shuttered-mode Westinghouse 7290 vidicon required for the MacDonald camera. While 7038's have been used in modifications of the MacDonald camera, the low speed of the beam across the target area—of the order of 7.5 inches per second—results in very marginal performance of the vidicon. Therefore, it was decided to use the vidicon in the fast-scan mode and sample its output to synthesize slow-scan pictures.

table 2. The following sstv nets are in operation to promote slow-scan activity and help amateurs get started in sstv. The first two are primarily Canadian nets.

Tuesdays	2000 EST or EDT	3.740 MHz	Jack Barlow, VE3PW
Thursdays	2000 EST or EDT	14.180 MHz[1]	Syd Horne, VE3EGO
Saturdays	1900 GMT	14.230 MHz	Don Miller, W9NTP

1. listening for U. S. stations on 14.230 MHz.

The current sstv standards require the output to be equivalent to that from a 15-horizontal-line-per-second, 8-second-per-frame scanning raster. The easiest fast-scan raster to sample to obtain this output is 15 frames per second with a vertical frequency of about 3 kHz.

By displaying the fast-scan picture on an oscilloscope, rapid adjustment of electrical and optical focus, lighting and framing is possible, where previously up to eight seconds had to elapse before you could see the results of such adjustments. This is a tremendous operating advantage. The video path to the oscilloscope requires about 300 kHz bandwidth, but this only goes to the scope and a short length of shielded wire is adequate.

The fast-scan mode also permits the incorporation of automatic target control (atc) which automatically adjusts the target potential of the vidicon to compensate for changes in subject illumination. The time constants required to incorporate this in a slow-scan camera would render it useless, but with a 15 frames-per-second raster on the vidicon, compensation is achieved in approximately one second for a change of illumination greater than 100:1. The operating advantages of this feature are obvious.

* Earlier the Canadian Telecommunications Bureau had been requested to authorize Canadian amateurs to conduct slow-scan television experiments in portions of the high-frequency bands available to the amateur service. Following a study of viewpoints expressed by various groups, an amendment was made to the General Radio Regulations Part II under the Radio Act to authorize special experimentation. Under this authority, permission was granted to Canadian amateurs to experiment with sstv in the bands listed in **table 1**, subject to revision on March 31, 1969, the end of the current licensing year. The Telecommunications Bureau provided additional assistance by arranging to monitor and record the spectrum of a typical sstv transmission.

Preliminary tests with the camera operating into a ssb transmitter drew the designers' attention to the danger of allowing the frequency-modulated subcarrier to exceed 2400 Hz. The filters of the transmitter wouldn't pass the higher frequencies, and the apparent signal dropout caused highlights to register as black on the receiving monitor; signal dropout also interfered with line and frame synchronization. A circuit was designed to defect frequency excursions beyond 2300 Hz and give an audible alarm whenever this occurs. The reference 2300-Hz signal used in this circuit is available with a flick of a switch so that its accuracy can be checked periodically.

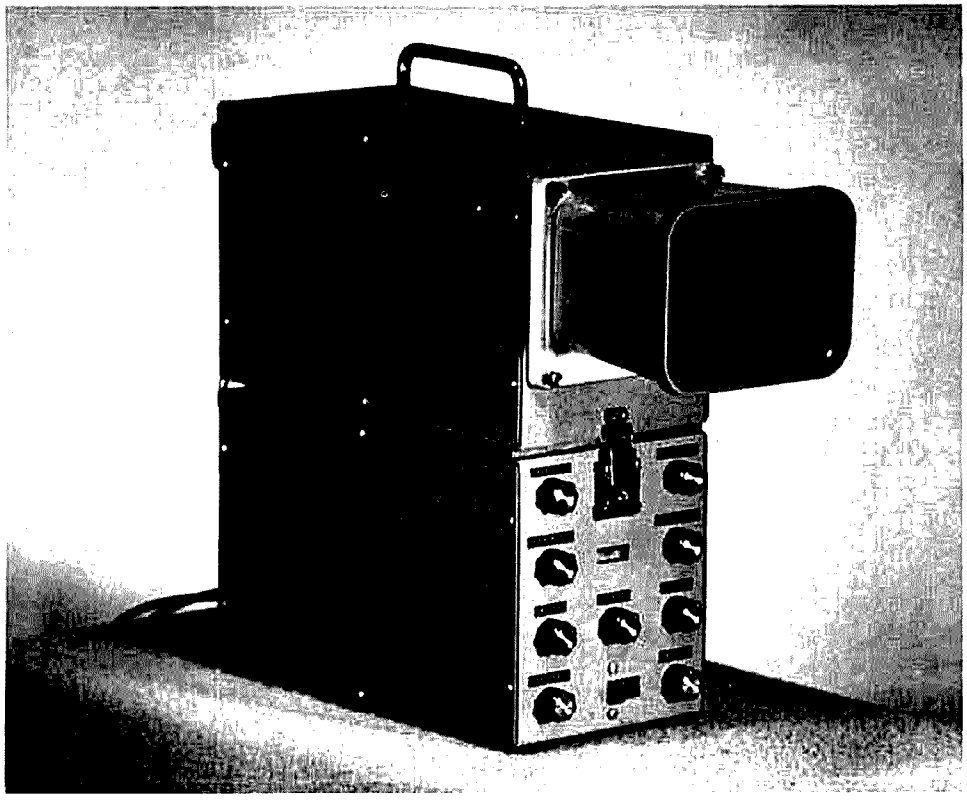
The camera consists of two units each $3^1/_2$x7x$8^1/_2$ inches. All dc voltages are regulated to minimize troubles due to hum and noise and line voltage changes. The only adjustment normally required in operation is the optical focusing of the lens.

the monitor

A magnetically focused and deflected five-inch cathode-ray tube with a P7 phosphor was selected for the monitor (5FP7). This carries a number of advantages for the solid-state approach as compared to a 5CP7A as used in the earlier monitor. The grid-cathode circuit of the 5FP7 can be, and usually is, at the grounded end of the high-voltage supply; this facilitates direct coupling which is almost a must for sstv work. The deflection coils are more easily driven by low-voltage solid-state circuitry than the deflec-

plates of an electro-static type tube. In addition, magnetically-focused tubes are characterized by a brighter image due to the geometry of the electron beam.

The specifications for the 5FP7 call for



Solid-state sampling slow-scan television monitor.

a minimum of 4000 and a maximum of 8000 volts dc anode voltage; 6000 volts were used in this design. In an older monitor, there was raster distortion whenever the line voltage changed (e.g., whenever the refrigerator upstairs turned on). There was also a certain amount of highlight aberration due to the increase in deflection sensitivity as the high voltage dropped because of the high beam current during such highlights. An electronically-regulated 6-kV supply that effectively cured these ailments was incorporated in the design.

table 3. Proposed slow-scan television standards.

Number of lines	**120**
Aspect ratio	**1:1**
Line frequency	**15 per sec**
Frame repetition rate	**1 per 8 sec**
Horizontal sync	**1200 Hz for 5 millisec**
Vertical sync	**1200 Hz for 30 millisec**
White frequency	**2300 Hz**
Black frequency	**1500 Hz**
Total bandwidth required	**1100 Hz**

Tuning an ssb receiver to provide intelligible reception of speech requires reasonably precise adjustment. The upward or downward shift of all frequencies due to slight mistuning is easily tolerated by the human ear. However, the reproduction of sstv images from the fm subcarrier requires the synch bursts to be within 50 Hz of 1200 Hz.

The fm demodulator designed for and used in this monitor was readily adapted to provide an output voltage proportional to the synch frequency. This is used to operate a tuning meter. A further step could be taken by controlling a variable reactance in the receiver circuitry to provide afc which would automatically correct for the slight frequency drift usually encountered.

The Tektronix polaroid viewing hood shown in the photo effectively eliminates reflection of light sources in the viewing area by the crt face and the yellow filter used to eliminate the blue flash of the P7 phosphor.

The monitor is housed in two utility cabinets hinged together at the back so that the top unit containing the 5FP7 can be tilted for optimum viewing angle. The over-all size of the monitor is 6x14x12 inches.

Unfortunately, it is impossible for the authors to provide more circuit or constructional particulars on the sampling camera and monitor until the proprietary designs of the Northern Electric Company are adequately protected. Hopefully they will be able to present the complete schematic and construction details as soon as the patent applications are filed.

Editor

summary

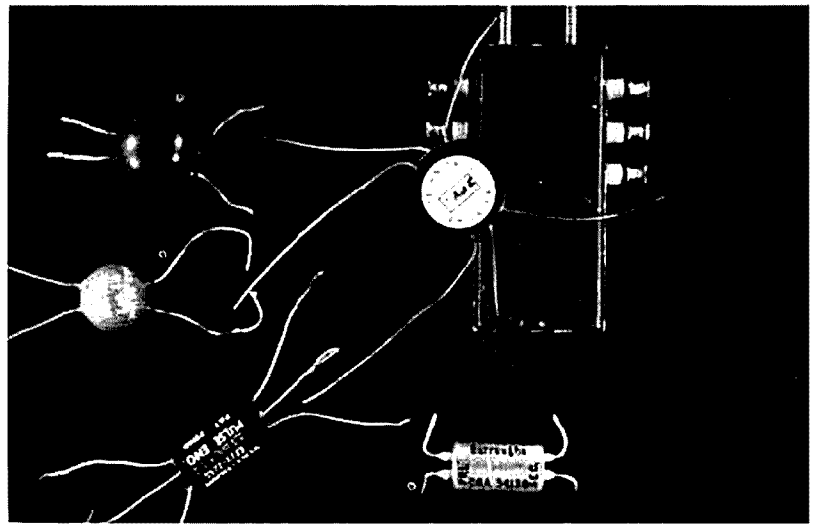
It is difficult to estimate the eventual impact of sstv on the amateur phone bands at the present time. However, a parallel can be drawn between the development and use of single sideband some 15 years ago and slow-scan television today. By overcoming many of the operating restrictions with the sstv sampling camera—along with the decided advantage of using readily available vidicons—it is reasonable to expect increased sstv activity in the future.

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ham radio

Pulse transformers for blocking oscillators.



enter the strange but useful blocking oscillator

Presenting some facts and figures on a versatile yet simple pulse generator

Hank Olson, W6GXN, P. O. Box 339, Menlo Park, California 94025

We are about to become acquainted with a singularly mischievous member of the oscillator family, whose apparent simplicity belies enough odd behavior to make the average electronic engineer weep all over his bagels.

The blocking oscillator is deliberately designed to produce nonsinusoidal waveforms; it usually produces pulses whose duration is short compared to pulse period (i.e., time between each pulse plus pulse length). While most oscillators are designed to have as low a feedback factor as possible (and yet remain self-starting), the blocking oscillator has a very large feedback factor. This, in combination with a low-Q tuned circuit, causes oscillations to build up rapidly, then die out for a time that's long compared to the pulse length (**fig. 1**). The transformer that is used with the blocking oscillator is not tuned except by stray capacitance, which usually assures the low-Q requirement.

how they work

Two forms of the blocking oscillator are in common use. The first is similar to the tuned-plate, tuned-grid or tickler-feed-back type oscillator. This is also shown in **fig. 1.** Of course, there is no "tuning," "plate," or "grid," so perhaps tickler-feedback is the better description. The second form is similar to the Hartley, as shown in **fig. 2.** Since only parasitic capacitance is used for tuning, the Colpitts form of circuit is not ordinarily used for blocking oscillators.

In **fig. 3** is a "real-life" blocking oscillator of the free-running, tickler-feedback type. This astable circuit produces pulses whose pulse length is primarily dependent upon the inductance and resistance of the collector winding of the transformer, and whose repetition rate is primarily dependent on the time constant, RC. A detailed formulation of the design factors and their interrelation is given in reference 1.

Table 1 is a listing of the pulse transformers, recommended for blocking oscillator service, made by one transformer manufacturer.[2] Note that pulse length increases with collector-winding inductance. Similar transformers are also manufactured by United Transformer Company and others. The turns ratio of the primary (collector) winding to secondary (base) winding is usually between 3 to 1 and 5 to 1. Linville[3] has shown that, for bipolar transistor blocking oscillators, a 5-to-1 ratio gives the sharpest pulses.

Since most amateurs are not in a position to buy specially constructed pulse transformers, they must either make them

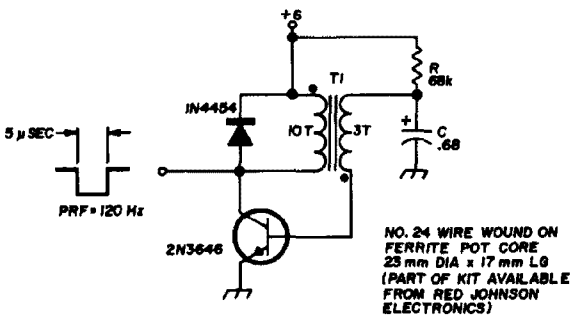


fig. 3. Blocking oscillator of the free-running tickler-feedback type.

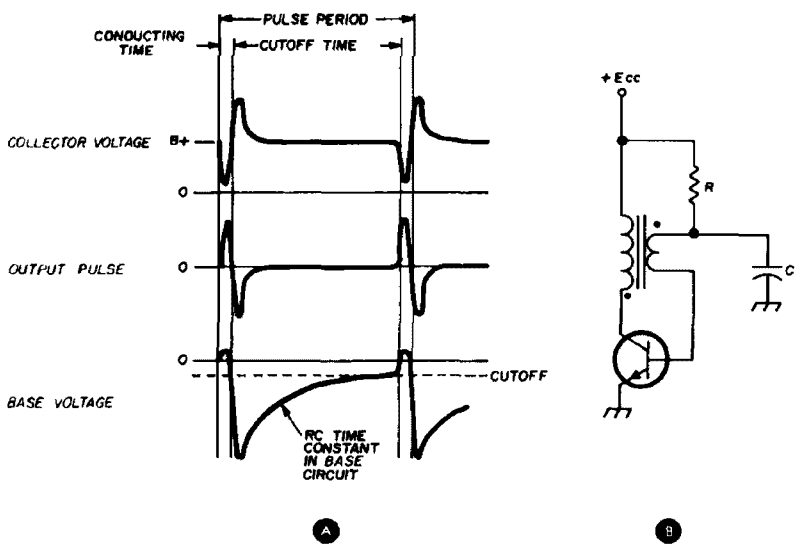


fig. 1. Single-swing blocking oscillator waveforms are shown in A. The circuit in B is a solid-state tickler-feedback blocking oscillator circuit.

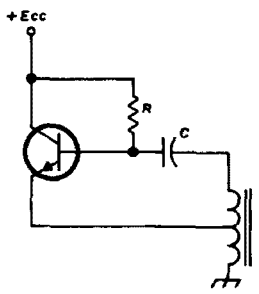


fig. 2. Hartley blocking oscillator circuit.

or use surplus types. However, rolling-your-own pulse transformer is really quite easy, now that ferrite cores are widely available. Because ferrite cores have such

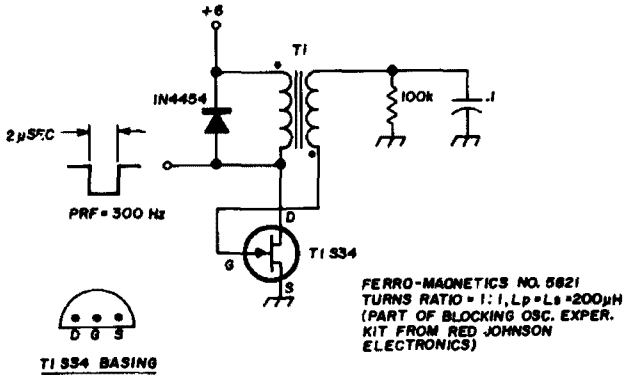


fig. 4. Blocking oscillator circuit that uses an fet with a tube-type pulse transformer.

high effective permeability, only a few hand-wound turns on a non-gapped core will provide enough inductance to produce a quite useful pulse width. Almost any of the closed-magnetic-path ferrite cores are adequate for pulse transformers, including pot cores, toroids, C-cores, E-I cores and U-I cores. The pot core used in **fig. 3** is part of a "blocking oscillator experimenter kit." *

table 1. Pulse transformers for blocking oscillator service.

Pulse Eng. Catalog No.	Turns Ratio	Pulse Width (μs)	Primary Inductance (mH)
2445	4:1:2	0.07 to 0.2	0.05
2444	4:1:2	0.1 to 0.2	0.1
2443	4:1:2	0.1 to 0.5	0.2
2671	4:1:4	0.1 to 0.5	0.2
2442	4:1:2	0.2 to 4.0	0.4
2672	4:1:4	0.2 to 4.0	0.4
2248	4:1:2	1 to 12	2.5
2673	4:1:4	1 to 12	3.0
2249	4:1:2	1 to 20	5.0
2250	4:1:2	1 to 50	10.0

* Available for $2.00 from Red Johnson Electronics, 440 Pepper Avenue, Palo Alto, California 94306.

Most of the pulse transformers available in surplus are for the older tube-type blocking oscillators. The photograph shows a selection of these that were available in my local area surplus stores. The physical form of such small transformers seems to be quite distinct, and if you find one of these "gum balls," it is pretty likely to be a small pulse transformer.

The surplus tube-type pulse transformers **can** be used directly in bipolar transistor blocking oscillators, but the 1:1 ratio that most offer is too high to provide the best-looking pulses. But they can also be used

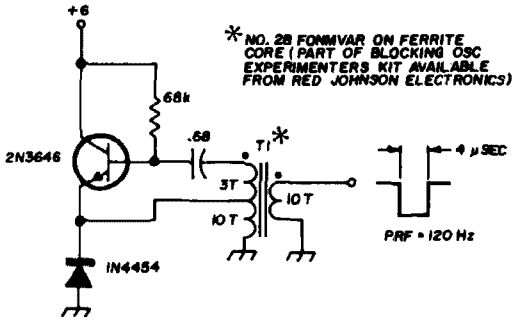


fig. 5. Hartley-type blocking oscillator.

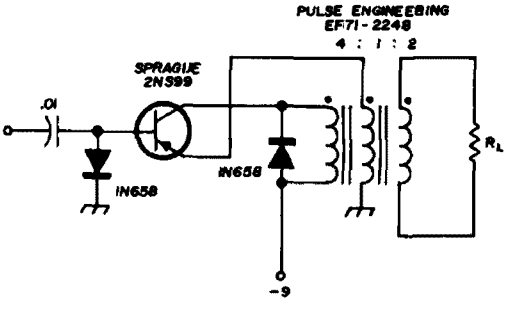
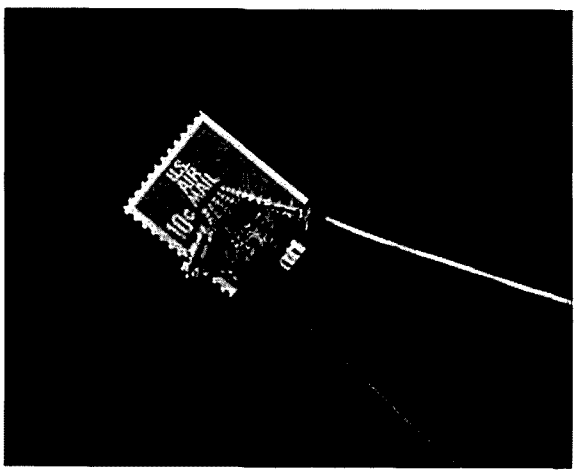


fig. 6. Blocking oscillator that is not free-running but is used as a one-shot.

in an fet blocking oscillator that is more nearly like the tube circuit for which they were designed, as in **fig. 4.**

When using an fet in a blocking oscillator it is best to use one that is de-

signed for switching, but such fet's are usually fairly expensive. The TIS34 (Texas Instruments) costs less than a dollar and has an $r_{ds(on)}$ of only about 200 ohms, and so this was the choice for **fig. 4.** The TIS34 is normally used for vhf rf amplifier service and has had wide acceptance in amateur two-meter service.



Miniature blocking oscillator using the circuit shown in fig. 7.

The circuits of **figs. 3** and **4** are the free-running tickler-feedback type. **Fig. 5** shows a Hartley-type blocking oscillator. **Fig. 6** shows a blocking oscillator that is not free-running, but used as a one-shot. That is, the blocking oscillator produces one pulse for each trigger input. The circuit of **fig. 6,** by the way, is that recommended by Pulse Engineering to obtain the pulse widths specified using their transformers.[2]

integrated circuit pulsers

For those who favor an integrated circuit to produce pulses, because the IC is smaller and outwardly simpler, **fig. 7** shows a circuit and picture of a blocking oscillator producing 100-ns pulses. Yes, friends, discrete circuits can be small, too. The ferrite core in this case is a ferrite shielding bead costing about 10c; the transistor is a Motorola plastic Micro-T at $3.00. (A 2N3904 could be substituted at less cost if the small size isn't required.)

A circuit related to the blocking oscillator is shown in **fig. 8.** This circuit achieves the phase shift required for oscillation by means of a delay line. The waveform produced is a nearly square wave, whose period is twice that of the

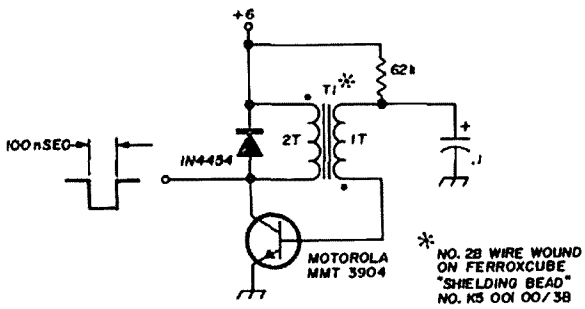


fig. 7. Miniature blocking oscillator circuit using a Motorola Micro-T transistor; completed circuit is shown in the photo.

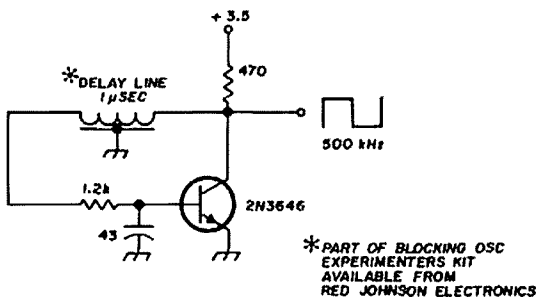


fig. 8. Circuit that achieves the phase shift necessary for oscillation from a delay line.

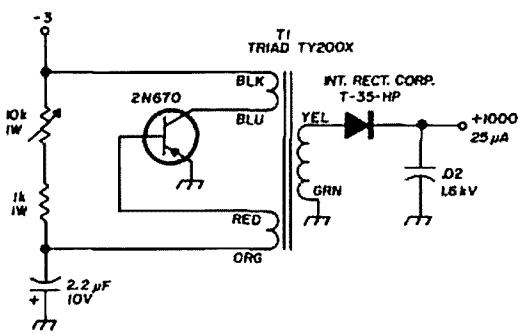


fig. 9. Blocking oscillator used in a high-voltage Geiger-counter power supply.

delay line. Since the ratio of pulse length to period is always fixed at 1/2, this circuit is not as versatile as the blocking oscillator. However, for sheer simplicity this oscillator has few peers and can serve as a clock for RTL systems.

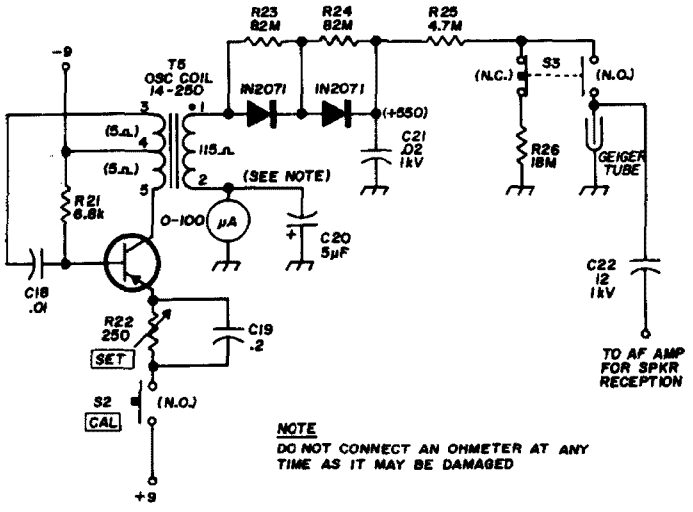


fig. 10. High-voltage Geiger-counter supply using a blocking oscillator.

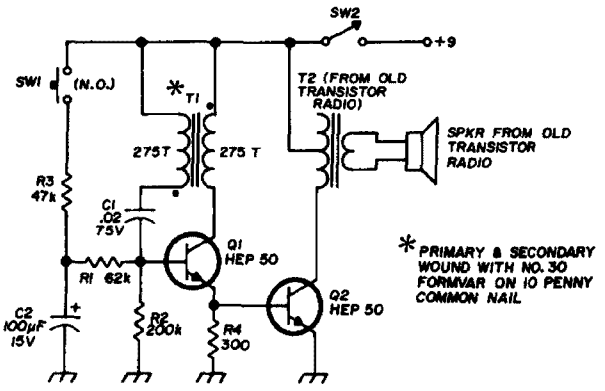
applications

Aside from the obvious use of the blocking oscillator to produce short pulses for use in electronic devices such as tv sets and radars, there are a number of other applications. One very important use of blocking oscillators is in dc-to-dc converters. Since the blocking oscillator has such a low-duty cycle (i.e., the pulse length is short compared to the period) it is particularly suitable for converters requiring high-voltage, low-current output. Geiger counters are typical users of such converters. A circuit is shown in **fig. 9.** Another similar circuit is shown in **fig. 10;** this circuit is available from John Meshna for about $5.00 as part of a surplus Geiger counter. Other blocking-oscillator-type dc-to-dc converters are contained in references 4, 5 and 6.

Another interesting use of the blocking oscillator is in electronic sirens. Such a circuit is shown in **fig. 11.**[7] When SW1 is closed, C2 charges through R3, increasing the available base voltage of a blocking oscillator. This increasing voltage increases the pulse rate of the blocking oscillator, while the pulse length stays more or less constant. When SW1 is opened C2 discharges through R1 and R2, causing the pulse rate to decrease again. The alternate, manual, closing and opening of SW1 produces a siren-like wail.

Other impulse-type services formerly using electromechanical contactors, such as electric auto fuel pumps and electric fence devices, are naturals for the blocking oscillator.[8] The substitution of a blocking oscillator for arcing contacts in these services reduces radio frequency interference and increases reliability. All-in-all, a very nice cousin of the standard oscillator.

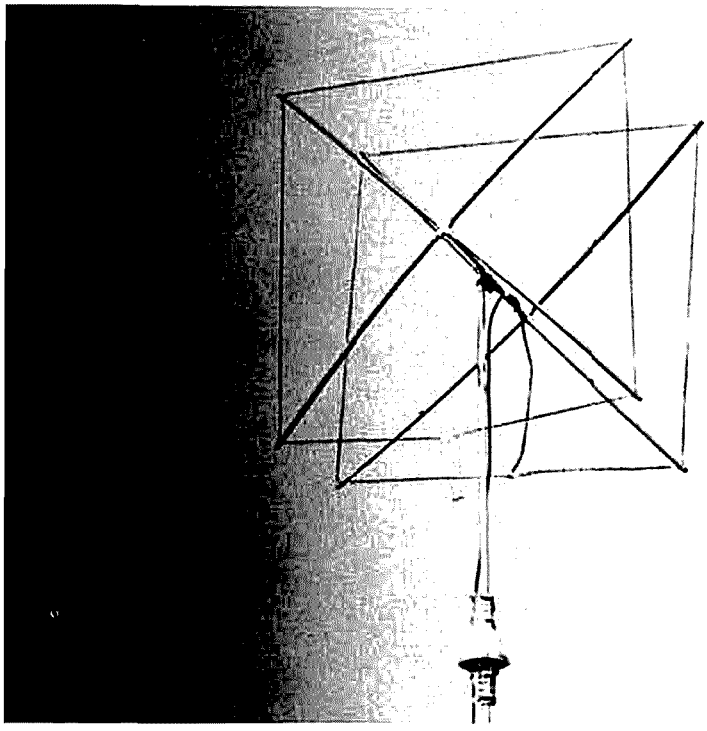
fig. 11. Electronic siren using a blocking oscillator; transistor Q2 is the audio amplifier.



references

1. "Handbook of Selected Semiconductor Circuits," *Buships NObsr 73231,* pp. 6-18 to 6-21 and 6-66 to 6-71.
2. *Pulse Engineering Catalog 206,* 1967, pp. 11-14.
3. J. Linville and R. Mattson, "Junction Transistor Blocking Oscillators," *Proceedings of the IRE,* November, 1955, pp. 1632-8.
4. "Transistors and Semiconductor Diodes," *RCA SCD-108A,* 1957, p. 31.
5. "Medium Power Germanium Transistor Power Supply Circuits," *Texas Instruments SC-1034,* October, 1960.
6. H. Olson, "A Self-Regulating DC to DC Converter." *73 Magazine,* May, 1966, p. 46.
7. "Hobby-Experimenter Projects Handbook," *Motorola HEP 400.*
8. "Motorola Semiconductor Circuits Manual," *Motorola,* 1964, pp. 7-31 through 7-34.

ham radio



an economy six-meter cubical quad

Here is a directional antenna you can build for one buck per dB of gain

Ev G. Taylor, W6DOR, 4100 Worthington Drive, North Highlands, California 95660

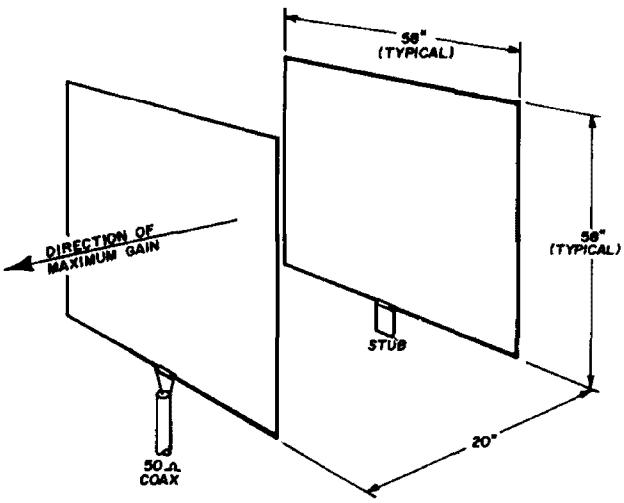
My first exposure to cubical quad antennas was right after World War II. A local amateur built one, and it worked very well. In fact he was doing as well as one of his ten-meter friends, who was using a three-element yagi. I built a quad also, then, but it didn't last long. I used a rather heavy wood and some dime-store shelf brackets; a Southern Indiana wind took care of it a few weeks after I put it up. Then I went back to yagis.

After several years and as many moves around the country, I finally settled in California, and I began to think seriously about the best antenna performance I could get for the least amount of money and labor. The quad won, and I must say Bill Orr's book[1] had a lot to do with it.

The six-meter quad I'm using now performs excellently and cost me about one

dollar per dB of gain. That's five dollars. If I use the gain figures you see in many ads for these antennas, then the cost would be something like fifty-eight cents per dB. That's because some of these ads base their claims on an isotropic reference source. This is engineering lanuage for an imaginary point source stuck somewhere out in space where there aren't any path losses and other earthbound aberrations. But I can say one thing for certain: my quad works.

fig. 1. Layout of the economy six-meter quad showing the major dimensions. Both the driven element and reflector are made the same length to start; the reflector is tuned with the shorted stub.



materials

The secret weapon that beats the high cost of materials for this antenna is rigid plastic connectors. Of course, you must realize the antenna is designed for six meters, which means that physical lengths are small compared to 20 meters, for example. But if this six-meter antenna were scaled up in size to 20 meters, the cost certainly would be competitive with any commercial quad on the market.

A list of material you'll need is given in **table 1.** It is possible to shave costs even more if you've got a fairly well-stocked junk box and are resourceful. For example, you can apply the time-honored technique of boiling a two-inch piece of wooden dowel in paraffin for an egg-insulator substitute, or you could even use a piece of plastic water pipe, drilled at each end. There's no reason you couldn't use a length of mop handle for the boom; however, you would have to devise some method of fastening it to the crossarms other than threading it. With a little ham ingenuity, you could easily cut the material costs by one-third. Would you believe a high-gain, directional antenna for $3.98 plus tax?

the crossarms

The material sources in **table 1** are suggestions of course. However, the plastic crossarms may be a little hard to find. They're made by Atlantic Research under the trademark G-S-R. My local source said they are rather uncommon and are made only by Atlantic Research, because patent rights haven't been issued to other plastics manufacturers. So if you can't locate any in your area, I've included the name and address of my source. These crossarms will accept ½-inch rigid plastic water pipe for the element supports, and the axial hole is threaded for steel pipe, which serves as the boom.

element supports

The plastic water pipe used for these parts comes in ten- and twenty-foot lengths. You'll need eight lengths, each 41 inches long. That's exactly 27 feet, 4 inches. I'd recommend buying one twenty-foot and two ten-foot lengths. Then you'll have a little extra material for insulators or other projects. This material is easy to work and has a good dielectric constant, at least up to the six-meter region.

elements

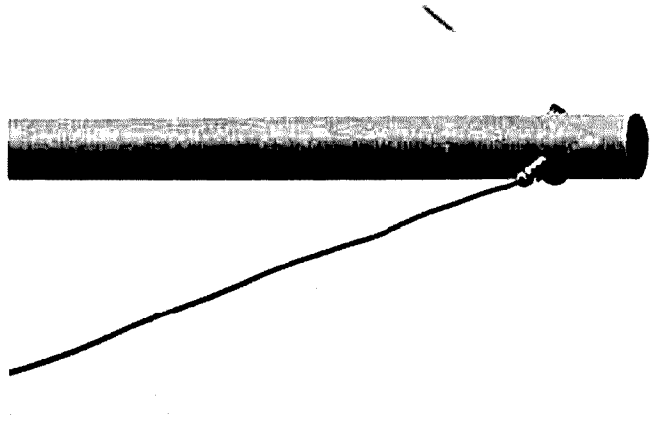
The sketch of **fig. 1** indicates that you'll need 38 feet, 8 inches of wire for driven element and reflector. Almost any type of wire can be used, but I recommend copper-clad steel. This wire is designed for antennas. It's a little more difficult to work with than pure copper wire, but it doesn't stretch. Another substitute is Sears' electric

fence wire. It costs about $5.50 for a half-mile length of number 18 galvanized. Element diameter doesn't affect antenna bandwidth in a quad to any great extent, because the Q of a closed-loop antenna system is low. I've shown fifty feet of wire as the recommended quantity in **table 1.** This allows a little extra for the stub and fastenings.

construction

Assemble boom and crossarms first. The steel pipe boom should be threaded at both ends. If you decide to use a mop handle boom, it should be easy to force fit the wood into the plastic crossarms I'm a great believer in epoxy. Although I didn't use a wooden boom, it should work all right if you use epoxy to weld boom to crossarms. It's a lot lighter in weight than the steel pipe, and it's free.

Drill holes for the wire in one end of each of the eight element supports. These holes should be located one inch from the end. Now insert the one-foot length of 3/4-inch diameter dowel into the element supports where they join to the crossarms. This gives added strength.



Attaching the wire elements to the support arms.

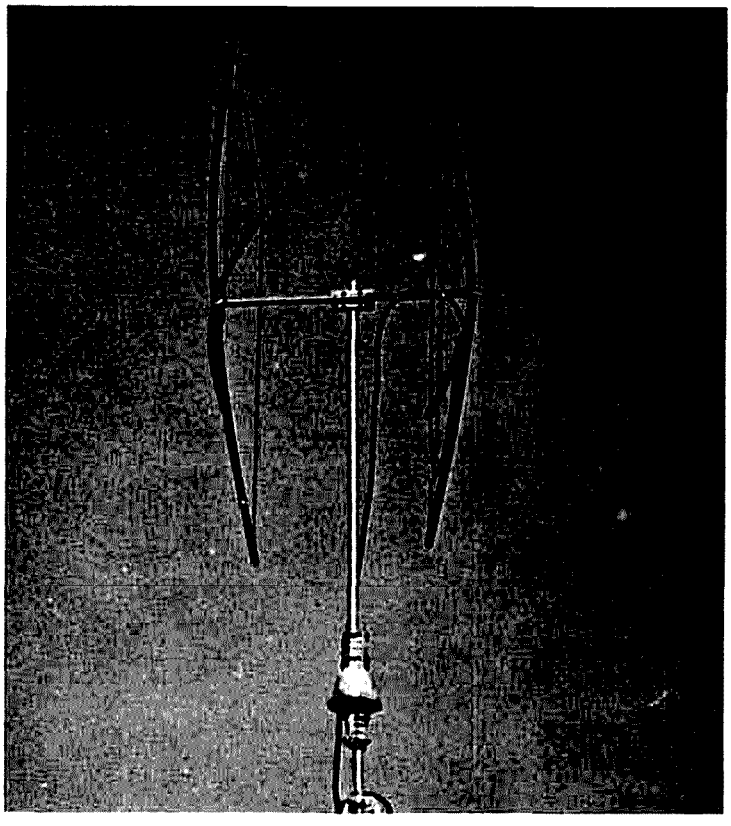
String the wire through the holes in the element supports and pull the wire up snugly. The element supports will then pull inward slightly. This adds to the stability

table 1. Material for the economy 6-meter quad.

part	material	quantity	suggested source or manufacturer
elements	no. 18 copperclad steel, or no. 12 AWG enamelled wire, hard drawn	50 feet	Junk box. (Alternate: Sears, Roebuck no. 18 electric fence wire; farm catalogue.)
elemont supports	plastic water pipe, 41-inch long by ½-inch	8 each	Lascopipe ½-inch PVC 1220 or equivalent.
crossarms	plastic water pipe "side outlet cross"	2 each	E/Tronics, 7300 Worthington Drive, North Highlands, California 95660 Part no. 1764-1; $4.00 for five crossarms.
boom	galvanized steel water pipe, 24-inch long by ½-inch diameter (see text)	1 each	Local hardware or plumbing supply (also see under "stiffener" below).
stiffener	wood dowel, one-foot long by 3/4-inch diameter	8 each	Mop handle, sanded and varnished (use marine spar varnish or opoxy).
hardware	egg insulators (2 each); Fahenstock clips (2 each); 6-32 cadmium-plated machine screw (1 each); flat washer (1 each); lock nut (1 each)		Local hardware supply (also see text).

of the whole affair. The assembly will then resemble a pair of back-to-back umbrellas.

Wrap six-inch lengths of wire around the elements as shown in the photo. An egg insulator serves as the termination for the driven element. The transmission line is soldered directly to the ends of the driven element. If you use coax cable, this means the center conductor and dielectric will be exposed to the elements. The best practice in a low-cost system like this is to give the exposed areas several coats of glyptal, Q-dope or any other good protective material.



side view of the economy quad.

There's no rf voltage at this point, but you'd be surprised what the transmission-line termination will look like in about six months if it's not protected. Be sure to allow adequate drying time between coats. Build up the protective coats gradually, then wrap the whole termination with Scotch number 33 electrical tape. Dope the tape thoroughly in the same manner.

Another egg insulator terminates the reflector, with the wire extending ten inches beyond the insulator. This is the stub, which makes up the five- or six-percent extra length required for this element. Short-circuit the stub with two Fahenstock clips connected back-to-back with a 6-32 cad-plated machine screw, flat washer and lock nut. Don't use dissimilar metals for this hardware, or you'll never be able to find the hardware when you decide to disassemble the antenna. Electrolytic action will deteriorate the parts, especially if you live near the coast.

tune-up

With the dimensions given in **fig. 1**, the antenna resonant frequency is 50.15 MHz. This was determined with a Measurements Corporation Megacycle Meter (grid dipper). Tune the stub, by pruning the wire, to about five percent lower in frequency than the driven element.

performance

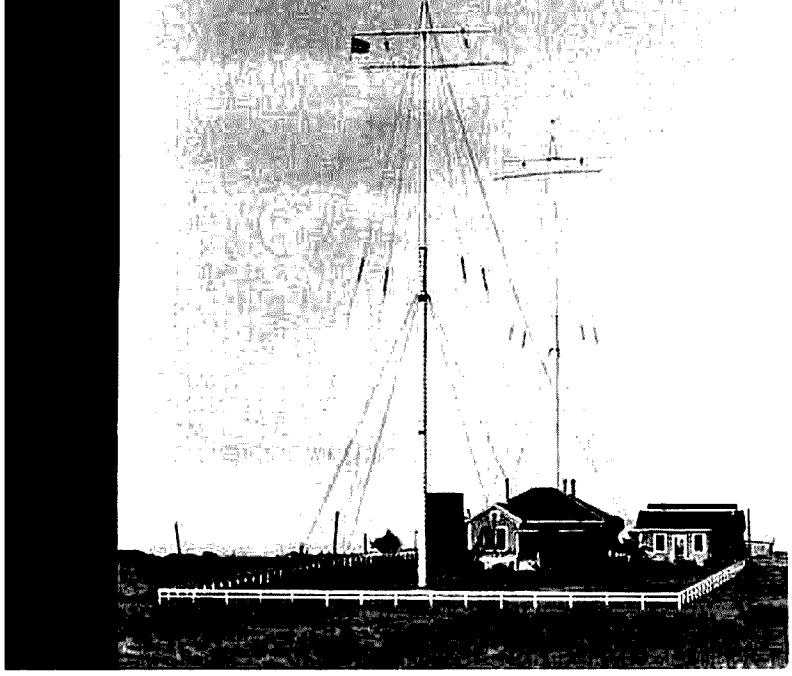
No antenna article seems to be complete without some report on performance, so I'll jump on the bandwagon with mine. Results to date have been quite good. A surprise opening in April 1968 netted 16 states in one weekend. With my Swan 250, contacts are being made almost every night with WA7GXM near Carson City, Nevada, with S-7 to S-9 reports. This is nearly 100 miles, over eight- and nine-thousand-foot peaks. Working the San Francisco bay area, 75 miles away, is a snap. And on April 1 and 29, 1968, I worked CE3QG during two F-layer openings with good results.

Perhaps a six-meter quad is the answer to the crosstown contacts that occur on twenty meters at the full legal power limit. Why support the power company? Come on up, the air's fine.

reference

1. Bill Orr, W6SAI, "All About Cubical Quads," Radio Publications, Wilton, Conn., 1959.

ham radio



Historical Collection, Union Title Insurance Company, San Diego, California.

wireless
point loma

Want to reminisce for a moment? Then listen to this old timer recall the early days of radio

Ed Marriner, W6BLZ, 528 Colima Street, La Jolla, California 92037

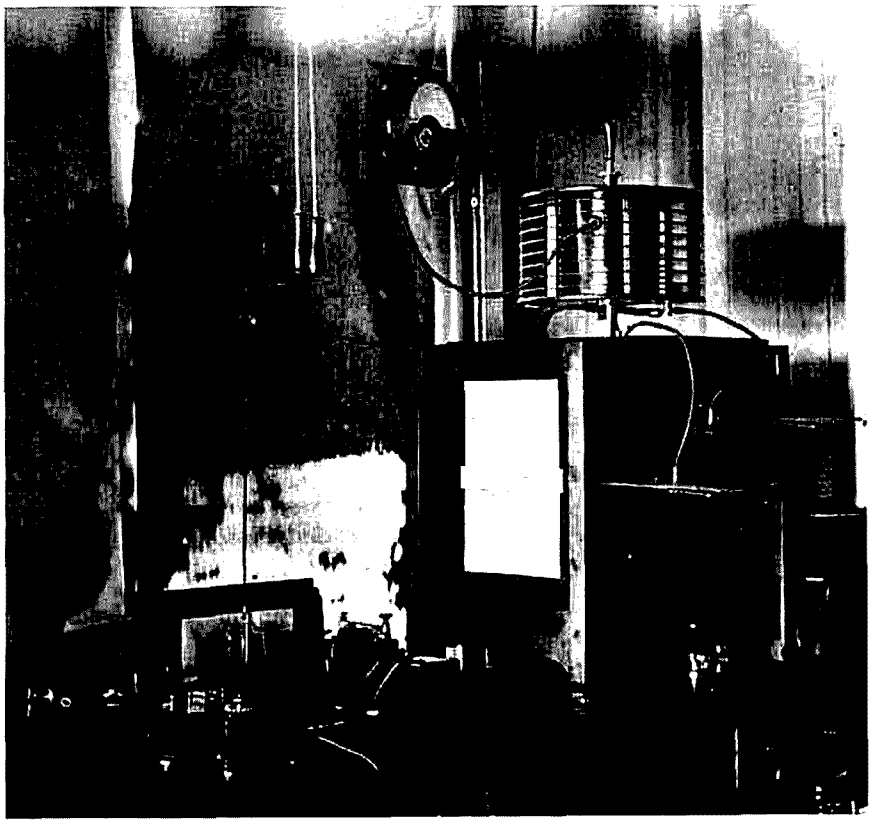
The fog lifted early that morning of 12 May 1906. A young Navy Chief Radioman, R. B. Stuart, unfurled a bright new flag and ran it up the mast of a new wireless station atop Point Loma, California. Then he went inside the yellow cottage and pressed a telegraph key. With an ear-splitting, 500-Hz crash (500 **cycles** in those days), a whitish-blue spark jumped across the ten-inch gap, and the U. S. Navy wireless was officially on the air.

This station was one of many established on the West Coast for the Navy during 1902-1903. The call sign, TL (later to be changed to NPL), was familiar to radio officers on ships hundreds of miles at sea as they steamed up the Mexican Coast from Cape Horn. (The Panama Canal was still a dream then—the first ship went through in August, 1914.)

The romance of those early days of maritime wireless was recalled for local San Diego amateurs as Chief Stuart, an engineer with the Eleventh Naval District for many years, told his story when the U. S. Navy Electronics Laboratory groundbreaking ceremony occurred at the old Point Loma wireless-station site on 24 June 1949.

The old radio shack where Chief Stuart first put TL on the air was now a transportation office. The old Massie spark transmitter was gone, as were some of the more advanced transmitters that superseded it in later years. In the weeds around the old site you could still see remains of angle-iron anchors where the old towers once stood.

All was now gone except for the memories of the whining notes of rotary-gap transmitters, not-so-whining notes of non-synchronous straight gaps, and the sweet, dc whistle of the old Poulsen arcs. Sound maudlin and corny? Perhaps. But let me tell some of you youngsters with your solid-state receivers and strip-line devices about these radio operators in the early days of wireless.



Point Loma station about 1905 with a coherer on the left. There's a hot-wire ammeter on the wall in back of the oscillation transformer; the glass-insulated transmitting capacitor is below.

the pioneers

A commercial radio operator working in the maritime service, from the turn of the century to World War II, was a fellow with a unique talent.

He could sit in a stuffy, five-by-nine-foot shack on an old rusty coal-burning tub and copy press at thirty words per minute on a beat-up Remington (provided by the owners if they felt like it; else he used his own). At the same time, he could carry on a conversation with the mate about their plans when the ship anchored in Manzanillo, while listening with one ear for his call sign on the 500-kHz guard band. (Split phones, you know.)

If they were in a heavy seaway, the old kettle would list fifteen degrees, and his bug would hang up on the dots, but Chief Stuart, at TL, couldn't tell the difference because the shipboard transmitter was being worked by a **radio operator.** It would have been great to use a mike and talk his message. Then he could have used that bug hand to pour his coffee and light his cigarette with the other. Yes, those fel-

lows were somewhat handicapped in those days. But they were radio operators . . .

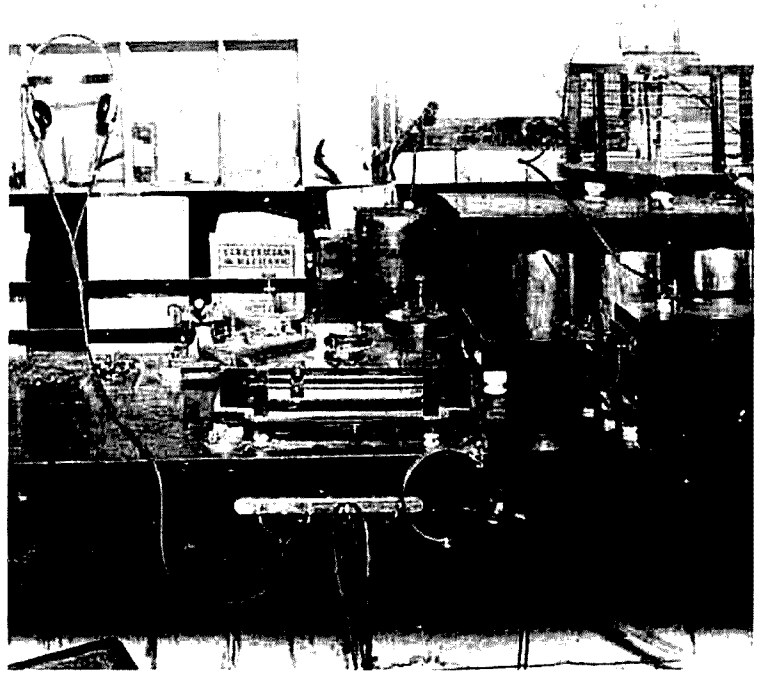
As I said, all was gone when Chief Stuart told his story, as were most of the operators. Si Slocum, W6JH, passed away a few years ago, as did Chief Stuart. A few of the early operators are still around, though. One is Fred Nickel, W6CKZ, and Admiral Ellery W. Stone, USNR (Ret.) who, as a young Lieutenant in the USNR, wrote "Elements of Radio Telegraphy."

early detectors

Across the room from Ellery, as he wrote the book, were the old copper helix and the marble-insulated switch panels. His desk was probably not much different from the earlier ones, which held the coherer detector, operated by tapping its glass tube with a doorbell clapper. This marvelous device would then uncohere the silver filings so the detector would receive again. It was a pretty slow business.

Other detectors were the magnetic detector, (which had to be rewound ever so often), the galena, silicon, or carborundum. After every message the galena would become desensitized because of the local large currents produced by the transmitter. The operator would then have to feel around on the surface of the crystal with the cat-whisker for another sensitive spot. You can well imagine the experience of the midwatch operator, who had just arrived at the station, after winding his way up the muddy canyon road in a horse and buggy. He'd probably been on liberty down in the small community of Roseville where he'd had a few beers. He was shaking so badly from the effects of the beer and cold that he wasn't in any condition

In the early days, amateurs often worked Naval and commercial stations. Gene Skinner used this station in San Diego in 1909 under the call "ES" to contact Navy ships in Magdalena Bay 500 miles away.



Early Naval stations on the West Coast

NPA	Cordova, Alaska
NPB	Sitka, Alaska
NPC	Bremerton, Washington
NPD	Tatoosh Island, Washington
NPE	North Head, Washington
NPF	Cape Blanco, Oregon
NPG	Mare Island, California
NPH	Vladivostok, Siberia (temporary)
NPI	Farallon Islands, California
NPJ	Balboa, Canal Zone
NPK	Pt. Arguello, California
NPL	Pt. Loma, California
NPM	Pearl Harbor, Hawaii
NPN	Guam
NPO	Cavito, Philippines
NPP	Peking, China
NPQ	St. Paul, Alaska
NPR	Dutch Harbor, Alaska
NPS	Kodiak Island, Alaska
NPT	Olongapo, Philippines
NPU	Tutuila, Samoa
NPV	Seward, Alaska
NPW	Eureka, California
NPX	San Pedro, California
NPY	St. George, Alaska
NPZ	Puget Sound, Washington

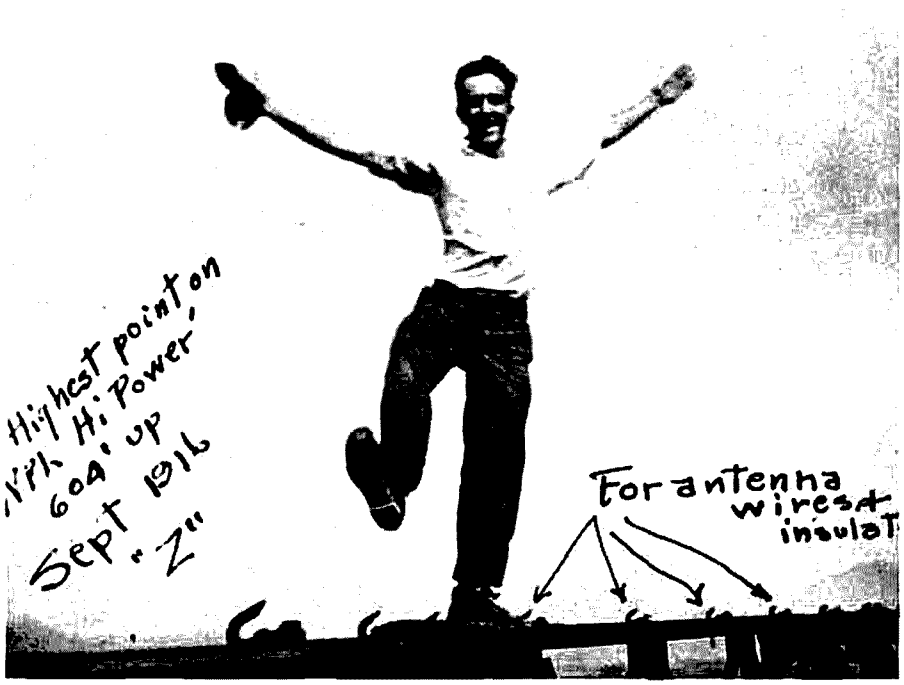
The call letters of Naval stations at yet an earlier period were:

TA	Cape Blanco, Oregon
TD	Tatoosh Island, Washington
TE	North Head, Oregon
TG	Mare Island, California
TH	Farallon Islands, California
TI	Goat Island, California
TJ	Pt. Arguello, California
TL (also TM)	Pt. Loma, California

to fuss around with any cat-whisker.

In those days the operators at TL did their own cooking, and many local residents lost chickens to night foraging parties from the station. In the daytime, the rabbit population of the Point would bounce out of sight as the operators with their shotguns approached, almost in the exact spot where Richard Henry Dana and a companion had hiked up over the point eighty years before.

Diego paper to the ships at sea, with "CQ de NPL-Press." Wireless operators on oil tankers, passenger ships and cruisers adjusted their cat-whiskers and began to copy the messages to put them in touch with the world. Point Loma's voice was available for both civilian and Navy traffic until 1915, when a German Telefunken transmitter was installed, and the old spark gap was silenced for all time, only to linger in the memories of many old-time



Joe Hallock, W7YA, Portland's now-retired FCC inspector standing on top of the 604-foot NPL tower in 1916.

the great white fleet

In 1908, NPL took part in wireless experiments with the Great White Fleet, which was sent around the world by Teddy Roosevelt. The Fleet had been working its way up the coast of Mexico to the coaling station at Magdalena Bay in lower California, and local amateurs competed with TL in contacting the ships. Dr. Lee De Forest was on board one of the ships, the USS Connecticut, the flagship. He had his audion-tube receiver aboard.

These early experiments with the USS Connecticut were soon forgotten as the station later was busy with its growing traffic, or was sending news from the San

wireless hams around San Diego who had hiked out to the station to watch.

Mr. Eugene Skinner remembers riding out to the station in 1909 on his motorbike after working TL on his spark rig. A couple of other early amateurs who used to contact TL were Harry Gough, K6AJ, and the late Carl Rogatsky, W6PG. These memories will probably bring letters from many of the old ops who listened to the sweet, 500-cycle note blasted through the air from the old Telefunken transmitter, which was installed in 1915 in Chollas Heights, east of San Diego.

ham radio

universal flea-power transmitter

Robert M. Brown, K2ZSQ, 5611 Middaugh Avenue, Downers Grove, Illinois 60515

With QRP societies claiming record memberships and ham magazines reporting flea-power DX over increasingly long expanses, many amateurs are tinkering with low-power circuits. The basic problem with most published designs is that the oscillator circuits are crucially dependent upon the L/C factors and related components.

The result is a circuit that will produce outstanding performance on 80 meters, but simply refuses to oscillate—without major circuit overhaul—on 160 or 20. The universal circuit illustrated in **fig. 1** overcomes this long-standing obstacle to true amateur-band flexibility.

five bands

By creating a parallel resonance, fundamental mode, a simple transistor oscillator can be built which will perform realiably over the range from 1 to 21.5 MHz. If you add tuned circuitry at the frequency desired, you can be on any ham band in that range.

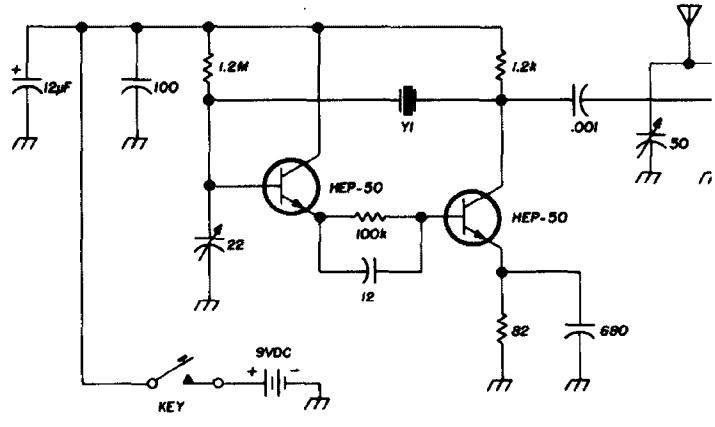
By using an extremely simple oscillator —with **no** tuning circuits—frequency drift due to aging and external ambient variations has been eliminated. The 22-pF variable stabilizes initial frequency shifts; it's a preset adjustment and doesn't have to be accessable from the front panel.

Similar in design to a Pierce oscillator with an emitter-follower inserted between the base of the second transistor and the crystal, the wide-band approach increases input impedance at the crystal input to the transistor. This impedance is normally quite low when compared with crystal characteristics. As frequency increases, the two emitter bias circuits aid in reducing gain attenuation.

The only way this wide-band oscillator can be upset from the stability standpoint is if the battery supply voltage fluctuates (you can easily cure any problems here by inserting a zener diode).

Referring to **fig. 1,** note that the output

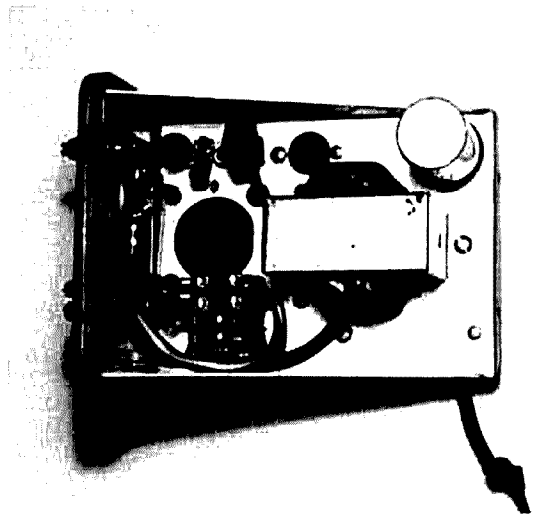
fig. 1. Circuit diagram for a flea-power transmitter for 160 through 20 meters.



circuit is tuned to 14 MHz. This is included here only for practical purposes. The tiny transmitter can be put on the air directly from this diagram—or you can adapt the output for any band you want.

Note that the transmitter is dependent on **fundamental-frequency** crystal oscillation. Doubling or tripling action is not recommended, unless a more elaborate output stage is added.

ham radio



the ideal t-r switch

You can modify any t-r switch to provide instant break-in and best signal reception at reasonable cost

Al Brogdon, K3KMO, RD 1, Box 390A, State College, Pennsylvania 16801

Two methods are in general use for providing an antenna in an amateur station installation. One is to use separate antennas for transmitting and receiving. Most amateurs interested in top receiver performance rule out this possibility, preferring instead to have the advantage of gain and directivity of their transmitting antenna for receiving. The other method is to use a single antenna for receiving and transmitting, and switch the antenna between receiver and transmitter with a relay or electronic transmit-receive (t-r) switch.

Both methods have advantages and disadvantages. For the cw operator, a t-r switch allows full break-in operation, which is necessary for traffic, net, DX and contest operating. If you've ever used full break-in, you can appreciate how nice it is to be able to monitor what's going on in the way of competition as you listen between code characters. But the t-r switch has one drawback. No t-r switch is as efficient as a direct connection between antenna and receiver. Many operators, therefore, compromise full break-in capability in favor of the best possible weak-signal performance and use an antenna changeover relay.

In this article I've presented the modification of a commercially available t-r switch to combine the advantages of elec-

tronic switching with those of an antenna changeover relay. The technique can be used with any commercial or homebrew t-r switch.

When I was faced with this choice, I wanted to have my cake and eat it, as usual, so I decided to combine the antenna relay with the t-r switch for the ideal combination of antenna changeover. The resulting improved t-r switch provides the capability of full break-in plus allowing direct connection of the antenna to the receiver.

the same, with only the mechanical considerations being different.

the change-over relay

The antenna changeover relay I used was a Potter and Brumfield KT11A, for 115 Vac operation. This dpdt relay has performed perfectly since its installation. One pole is used to switch the unbalanced coaxial feedline, and the other controls the transmitter. This provides a safety feature, because the transmitter can't be turned on until the relay removes the re-

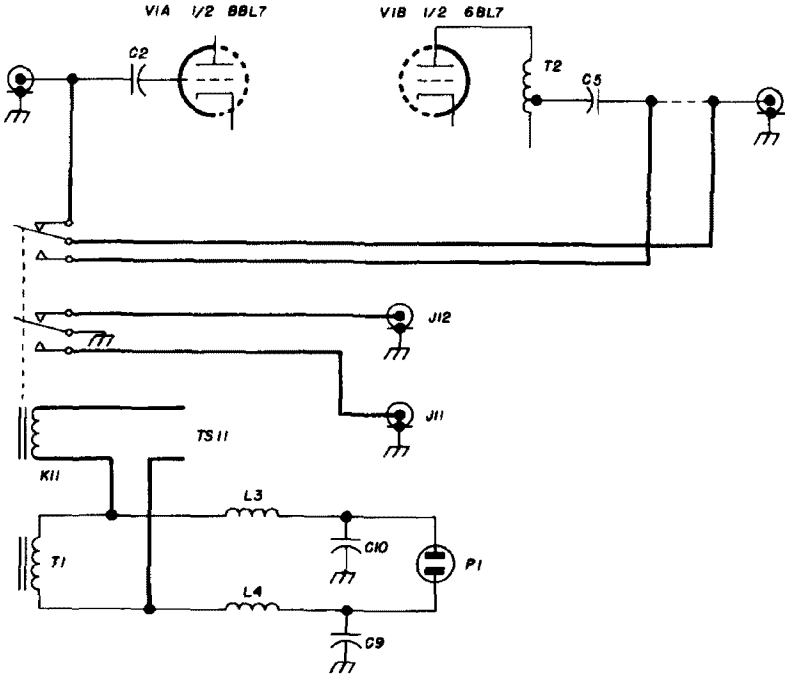


fig. 1. Modifications to the Johnson t-r switch are shown by the heavy lines. Relay K11 is Potter & Brumfield KT11A (115 Vac).

the t-r switch

The switch I used for this application was the Johnson model 250-39 t-r switch. It is compact, reliable, and inexpensive and performs well. Furthermore, exactly the right amount of vacant space is available in the case to accommodate the antenna changeover relay, and just the right amount of space is on the front panel for the required new connectors.

I refer to this particular t-r switch in the following discussion, but any t-r switch, either commercial or homebrew, can be used. The electrical connections will be

ceiver input from the antenna feedline. The transmit switch operates the relay, one pole of which turns on the transmitter B plus.

construction

The changes to the Johnson t-r switch to add the relay are shown in **fig. 1,** with the modifications indicated by heavy lines. TS11 is a bakelite terminal strip, and J11 and J12 are RCA phono connectors. These carry control circuits, and any suitable substitute can be used. An spst switch connected to the terminals of TS11 is the

transmit switch. J11 provides the ground-to-transmit function to activate the transmitter.

If the transmit control for your transmitter requires this line to be off ground, then another terminal strip or an ungrounded connector can be used to carry this control line. J12 provides a ground on the receive function, which may be used to mute the receiver for phone operation.

connections to the station

To connect the t-r switch into your station, make the three coax connections (to receiver, transmitter, and antenna) as indicated on the t-r switch front panel. Connect the station transmit control switch to the terminals of TS11. Connect J11 and J12 as described for transmitter and receiver control. It's that simple; you're now ready to go.

operation

With the station-control switch in the standby position, the antenna is relay-connected to the receiver, and the transmitter is switched to standby by the same relay. This provides best weak-signal reception. When the station-control switch is in the transmit position, the receiver is disconnected from the antenna and connected to the t-r switch output. At the same time, the transmitter is relayed on, ready for full break-in operation.

The photo shows the inside of the t-r switch with the added relay and connectors. It makes a very neat package. If you prefer not to modify your t-r switch, an external antenna changeover relay can be wired to provide the same switching action. A Dow-Key model DK60-2C can be used, with the relay's auxiliary contacts wired for transmitter control in the manner described.

The cost of the relay and connectors for modifying the t-r switch comes to about $4.25. The complete t-r switch described here can be duplicated as shown for a little less than $35. It's a low price to pay for the versatility and good performance obtained.

ham radio

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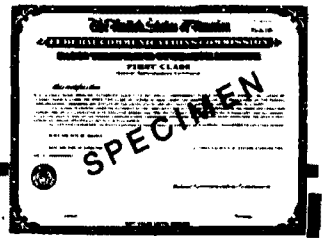
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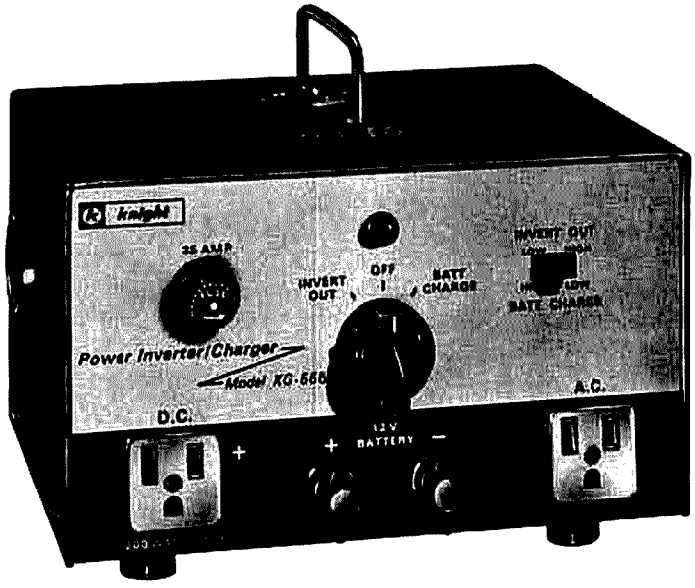
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knight-kit
power inverter/charger

Jim Fisk, W1DTY, Box 25, Rindge, N.H. 03461

Most amateurs, when they want to operate mobile or portable, buy one of the accessory dc-to-dc converters that goes along with their transceiver. The price will vary all the way from seventy to two-hundred dollars, depending on the manufacturer and power output that is required. A dc-to-dc converter that is designed specifically for a piece of equipment is probably the best approach for **daily** mobile operation, but if you only mobile occasionally or operate portable only on field day, you probably are reluctant to spend the money for a dc-to-dc converter that more or less is limited to one particular piece of gear.

The KG-666 Power Inverter/Charger kit, manufactured by Knightkit, is the ideal answer to the amateur who only wants (or needs) to operate on battery power occasionally. When operated from a 12-volt battery, this unit will provide 200 watts of 110-Vac power. In addition, when you're not operating mobile or portable, the power inverter/charger can be used for powering soldering irons, drills or even television sets from a 12-volt battery; or it may be used as a battery charger with a self-regulating battery charge tapering from 6 amps. If 200 watts isn't enough to handle your requirement, two of the KG-666's may be connected in parallel. The ac output of the inverter is 55 to 65 hertz depending on load and battery voltage; the tapped secondary permits selection of 105 or 120 volts.

A simplified circuit diagram of the KG-666 is shown in **fig. 1**—the switch is in the inverter position. The switch across the transformer secondary selects either 105 or 120 Vac output. When the switch is turned on—with the unit connected as an inverter as shown here—the four 2N443 transistors are heavily forward biased by the 200-ohm resistor in the center tap of the transformer feedback winding. Since the transistors are not exactly matched, one of the parallel-connected sets will conduct slightly more than the other set, and will initiate oscillation. As soon as base current begins to flow through the feedback winding, the diode in the center tap of the feedback winding is forward biased, and reduces bias to a lower level.

The transistors are protected from switching transients by the 2-μF capacitor and the

two back-to-back diodes connected across the primary of the transformer. In the diode despiking circuit, the 50-μF capacitor is charged to the peak voltage across the winding, and the two diodes are reverse biased. Any transients that tend to exceed the voltage stored in the capacitor are short circuited through the diodes.

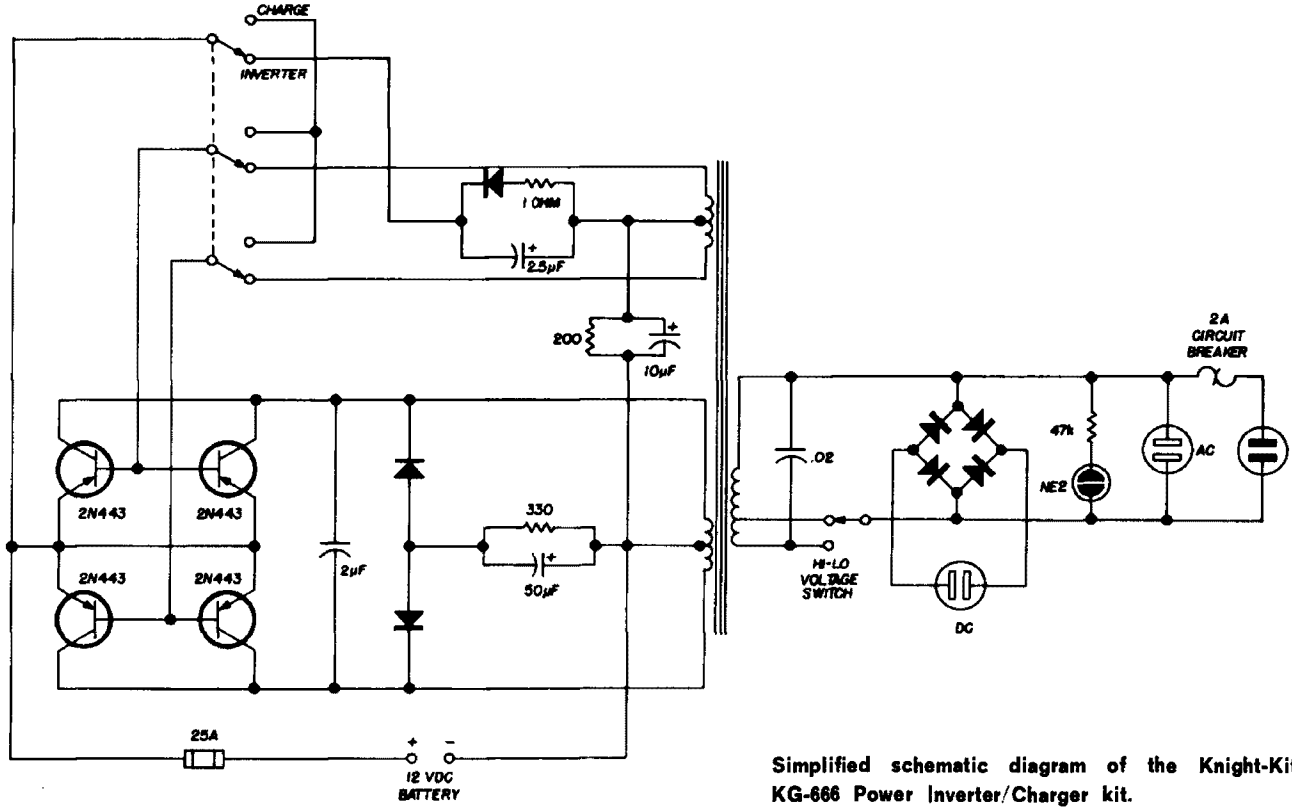
The square wave ac output of the inverter is converted to pulsating dc by the diode bridge. This output is useful for equipment that does not operate efficiently from square-wave ac such as the universal motors commonly found in electric shavers and inexpensive drill motors.

When the inverter/charge switch is pushed over to the charge position, the feedback winding is not in the circuit anymore; 115 to 120 volts is applied to the input through a 2-amp circuit breaker. When the KG-666 is used as a battery charger, the inverter transformer is used as a stepdown transformer to 12 Vac. The base of each transistor is connected to its emitter through the inverter/charge switch and two transistors are used as diodes in each leg of the full-wave rectifier circuit.

The KG-666 is extremely compact, measuring 8 inches wide, 4-3/4-inches high and 8$^1/_4$ inches deep. Total weight is 8 pounds. Battery connections are provided on the front panel through two large thumb-screw terminals. Two standard ac outlets on the front panel provide either 115 Vac or Vdc.

If you are interested in a utility emergency power supply for your shack, the KG-666 is worth looking into. It can be used to keep a large 12-Vdc automobile battery charged when 115-Vac power is available. If you should have a power failure, simply plug in your transceiver, transfer the switch to the inverter position and you can be on the air. At available full load—200 watts for the KG-666—the total load on your battery is about 20 amps. However, this assumes a constant load. With ssb or CW operation, the duty cycle is fairly low, so battery life should be relatively long. You should be able to estimate expected battery life depending on the transceiver you're using and the size and age of the storage battery.

ham radio



Simplified schematic diagram of the Knight-Kit KG-666 Power Inverter/Charger kit.

propagation predictions for april

Victor R. Frank, WB6KAP, 12450 Skyline Boulevard, Woodside, California 94062

During the past six months, amateurs in the Northern Hemisphere have enjoyed excellent daytime propagation on the higher bands (20, 15 and 10) and excellent nightime conditions on the lower bands (40, 80 and 160). Ten has been open almost every day; absorption has been low enough that 20-meter contacts could be made over the pole to stations in the Northern Hemisphere almost any time (muf's permitting) during quiet magnetic conditions; noise levels have been low... it was too good to last.

Seasonal changes begun last month should be complete by the end of this month. Accordingly, 160-meter maximum range predictions are being dropped in favor of 15 meters. Actually, a radio amateur with only modest equipment should be able to work anywhere in the world on 15 meters if the path is in sunlight and not within 30° of the North Pole. However, eventually even a 15-meter signal is absorbed, usually just after the first transit of the earth. The 15-meter range predictions are based on the same station parameters as 20 meters, 100 watts CW and a 6-dB gain antenna.

working through the twilight zone

The 15-meter band curves suggest the likelihood of round-the-world propagation (RTW), which may be more an annoyance than a useful propagation path. This is because some signal component with much less time delay is usually present, and the combination usually has at least two components with relative delays up to 1/7 second, causing difficult copy. The muf over the entire path does not have to exceed the operating frequency, but the muf at the end control points does.

If you wish to verify the existence of RTW you should first find yourself a partner—unless you have two beams and a receiver which will recover in much less than 100 milliseconds. The two stations

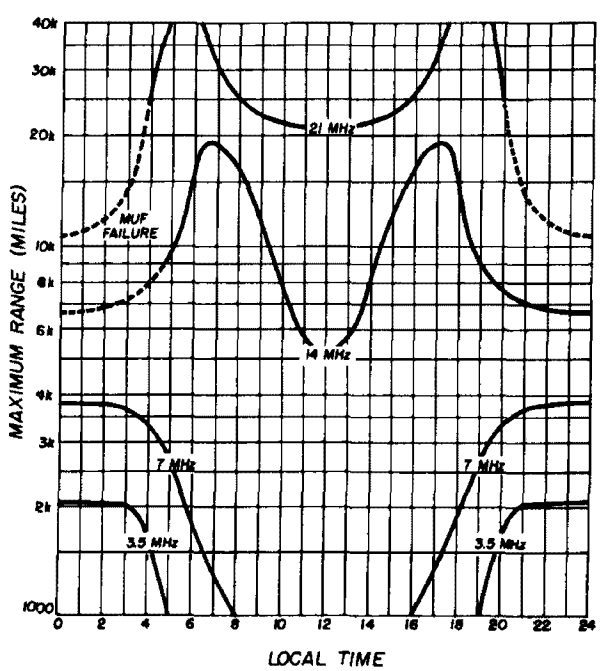


fig. 1. Maximum range vs local time for propagation to the north from mid-latitude United States (38° N).

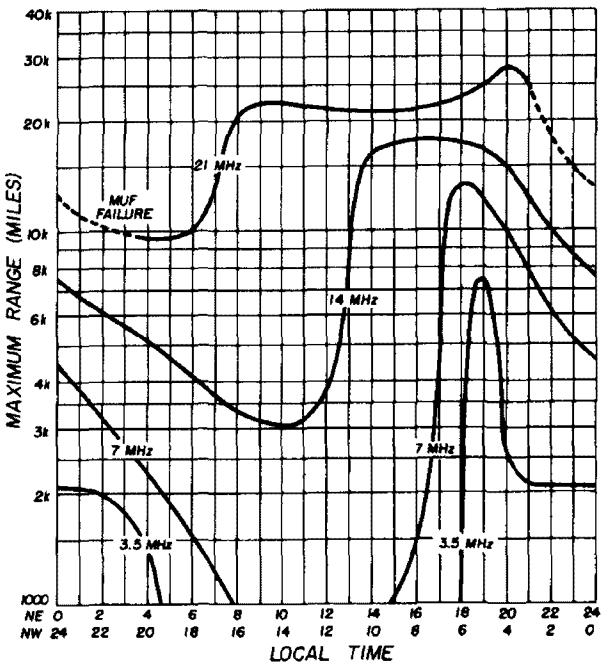


fig. 2. Maximum range vs local time for propagation to the northeast (top time scale) and the northwest (bottom time scale).

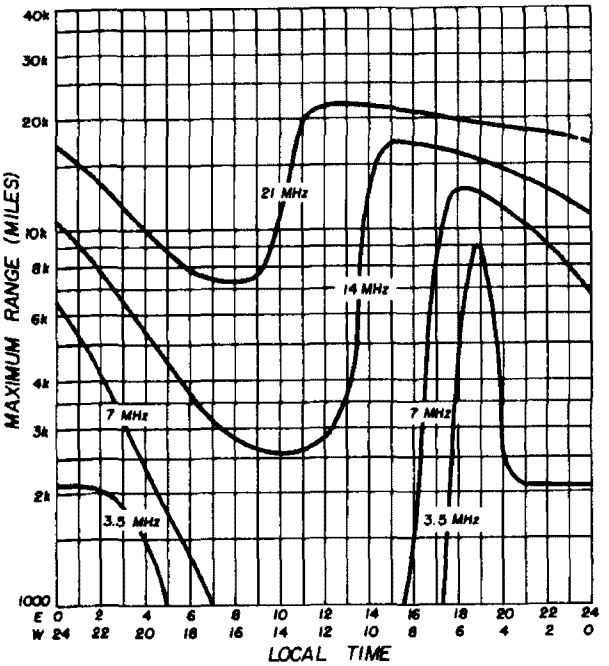


fig. 3. Maximum range vs local time for propagation to the east (top time scale) and the west (bottom time scale).

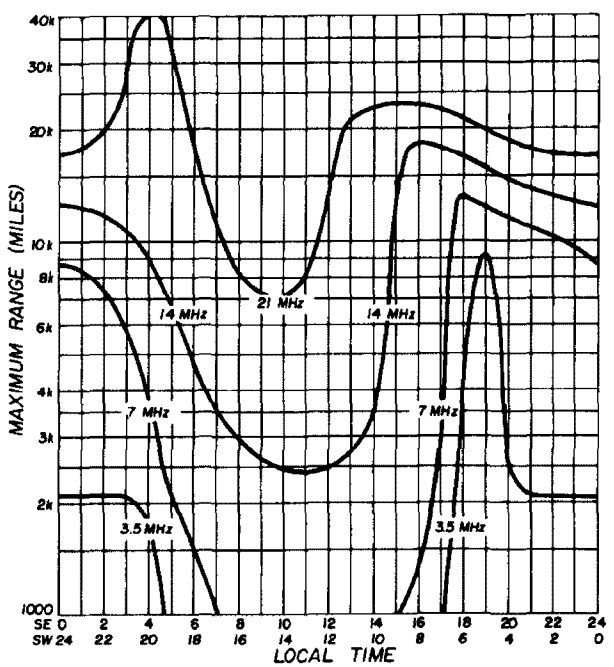


fig. 4. Maximum range vs local time for propagation to the southeast (top time scale) and the southwest (bottom time scale).

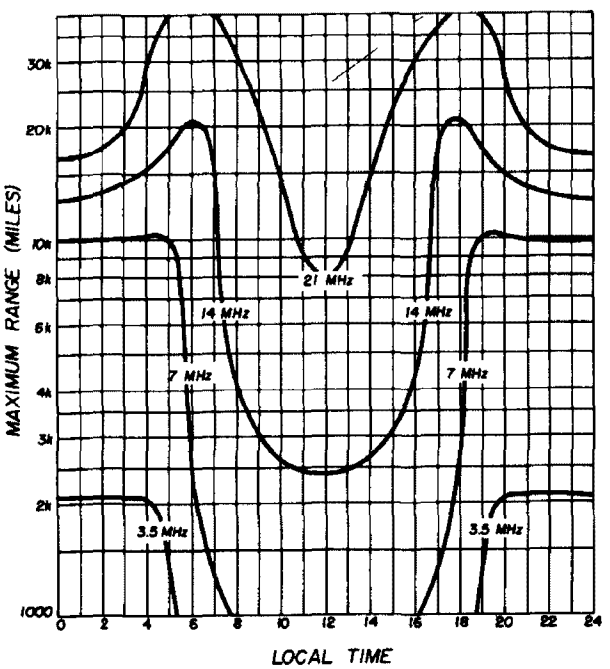


fig. 5. Maximum range vs local time for propagation to the south from 38° N latitude.

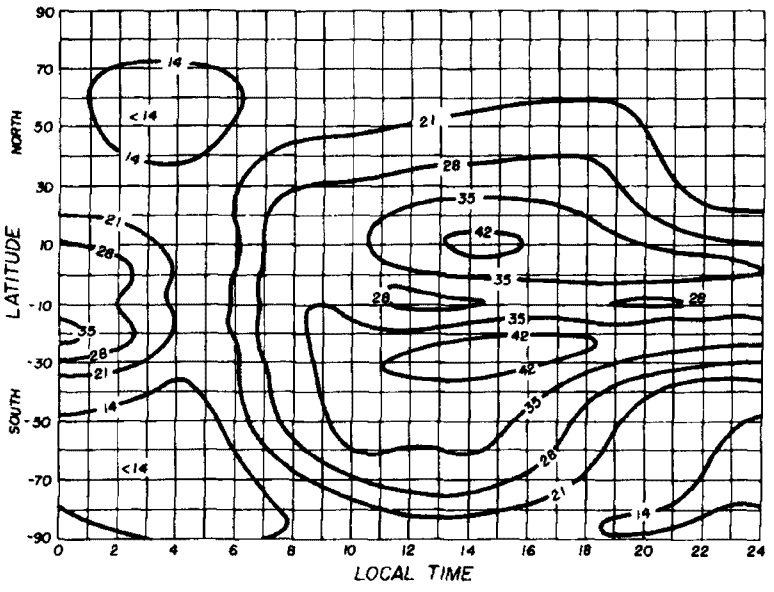


fig. 6. Time chart of predicted median muf for April 1969, centered on 90° W longitude.

ideally should be within each other's skip zones (600 miles). Pick a time from the maximum range charts for the maximum distance in the direction (and reverse direction) you intend to work. Best RTW conditions usually exist near the dawn/dusk line (NW-SW near sunset, NW-SE after sunrise). The positions of the twilight zone in the E-region is marked on each chart. The two stations should point their beams in opposite directions, and the station with the most power or noisiest location should transmit CW or ssb, and the other station should listen for an echo.

If the direct or scattered signal and the RTW signal are about the same amplitude, code copy at speeds much in excess of 15 wpm may be difficult. Listening to such a signal the first time is an eerie experience.

maximum usable frequencies

The predicted median F2-layer 4000-kilometer muf's for April 1969 are shown in the time chart of **fig. 6,** which is centered on 90° W. longitude. These muf's may be pessimistic if ionospheric soundings made last April at Point Arguello,

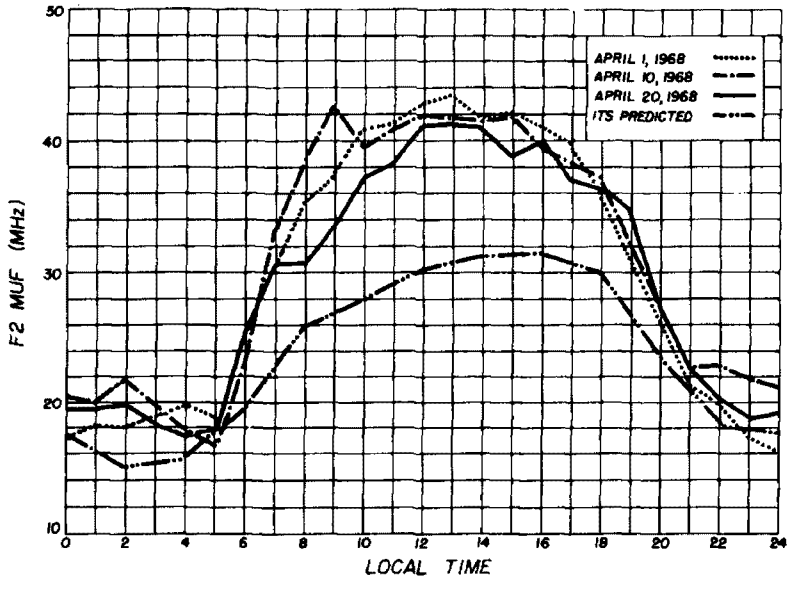


fig. 7. Muf vs time of day scaled from April 1968 vertical-incidence ionograms from Pt. Arguello, California.

California are typical of what can be expected this April. **Fig. 7** shows a comparison between muf's vs time of day scaled from Point Arguello ionograms on April 1, 10 and 20, 1968 and the median muf (4000) F2 predicted by ITS for the same point.

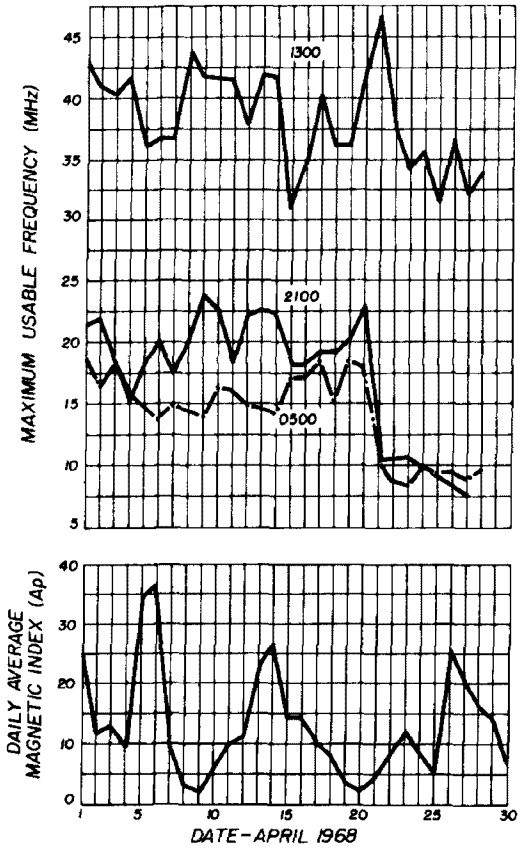


fig. 8. Muf vs day of month of April 1968 for hours of 0500, 1300 and 2100 local time scaled from Pt. Arguello ionograms.

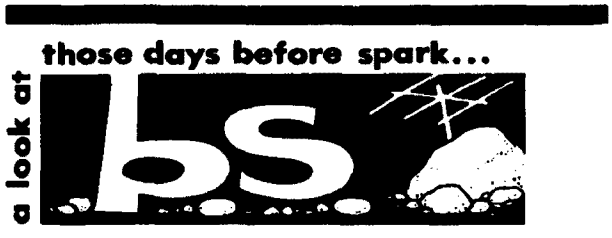
A better idea of conditions that may occur during the month may be gained from referring to **fig. 8;** this graph shows the muf's at 0500, 1300 and 2100 local time for each day of April 1968 along with the daily average magnetic index, Ap. Magnetic storms occurred on April 1; 1300, April 5 (gmt) to 1800, April 7; 0400, April 12 to 2100, April 14; and 2100, April 25 to 1100, April 28. Interestingly enough, the storm with the highest magnetic index wasn't the one that affected propagation the most. From April 21 to 28, the bottom dropped out of the nightime ionosphere. Sporadic-E and spread-F were prevalent. Vertical incidence critical frequencies were below 3 MHz during most of that period as early as 2100 local time.

I can't predict whether such a storm will occur again this April. However, ignoring disturbed days, two trends appear on **fig. 8.** The 1300 muf decreases throughout the month while the 2100 muf increases. Twenty meters will probably be open to somewhere 24 hours a day during quiet conditions.

vhf

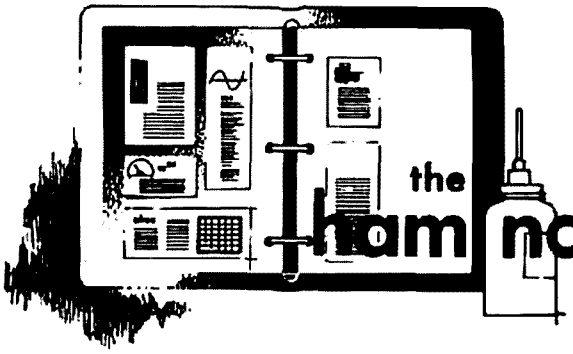
Transequatorial (TE) 50-MHz evening openings should continue through the month. Some sporadic-E is likely near the end of the month. The Lyrids meteor shower is expected between April 19 and April 23. Long distance tropospheric ducting across the Gulf of Mexico is predicted for two meters and higher frequencies.

ham radio



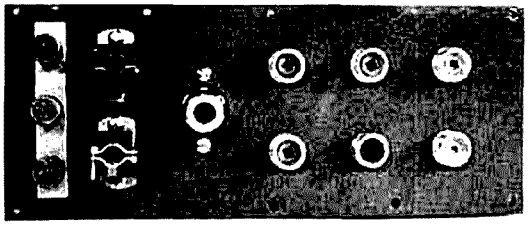


**"I triad mobile once . . .
yuh run into a lotta problems."**



coaxial feedthrough panel

Devising a simple and uncluttered method of bringing coax feedlines and rotor control cables into the shack is always a problem. After using a haywire lashup for years, I solved the problem neatly and efficiently with a small wall-mounted panel. Every wire except ac power is brought into the operating room through this panel.



The panel is made of 1/8-inch tempered masonite, 4 inches high by 9 1/2 inches wide. The coaxial feedthrough connectors are PL-258 straight adapters (Amphenol 83-1J) mounted snugly in a hole made with a 5/8-inch punch. Epoxy cement is applied around the connectors on the rear of the panel and will hold them in place forever if the epoxy is properly and thoroughly mixed.

The two Jones plugs which show in the photograph are used for control lines to antenna rotators; the small round plug is used for a remote swr meter. The metal strap is connected to a ground rod outside the house.

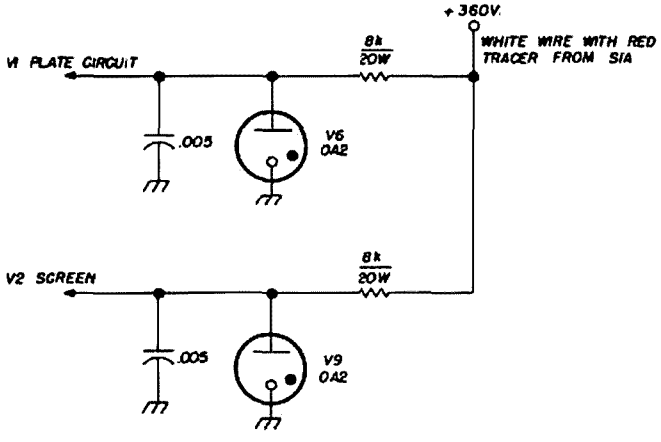
The panel may be painted with a spray can to match the decor or equipment. It is screwed to the wall over a 3 x 8 1/2-inch cutout in the wall near the radio desk; the coaxial cable and rotator control leads pass through the wall partitions to the outside. This panel is rather elaborate since it has six antenna lines and two rotor lines, but a smaller one for a simpler setup could be installed over a standard ac wall outlet box.

Edwin W. Hill, W3URE

T150A frequency stability

If your Knight T150A transmitter is chirping or showing other signs of instability, look for trouble in the voltage regulation circuit. The original circuit uses a 4000-ohm voltage-dropping resistor for the OA2 VR tube; this allows excessive current through the OA2 and causes early breakdown, removing regulation from the vfo and following stage. In my unit there was a five-volt variation!

To correct the problem, remove the 4000-ohm, 20-watt resistor (R37) from its mounting hole and install a 7-pin miniature socket in its place for an additonal VR tube (V9). Mount two 8000-ohm resistors vertically on top of the chassis near the VR tube socket and drill some holes in the chassis to accommodate wiring to them.



Connect the white wire with red tracer (formerly connected to R37) to a new terminal strip mounted under the chassis; run leads from this point to the two new resistors on top of the chassis. Now wire the two VR tubes as shown in **fig. 1.** Be sure to install another .005 μF disc capacitor across V9. With these simple changes, you should have no further troubles with instability.

Al Mumby, WB2MCP

dual-voltage power supply

Here is a regulated plus-and-minus power supply that uses a National Semiconductor LM-300 integrated-circuit voltage regulator. By changing the values of R2 as shown in **fig. 2,** the output voltages are ±10, ±12 or ±15 volts. The transformer charges the input capacitors to nearly 25 volts at light loads, so there is no problem in obtaining 15 volts output. The positive supply is regulated by the LM-300 while the negative supply is "slaved" to the positive one. The slaved supply will follow the positive supply from 4 to 15 volts, but maximum output current is less at higher output voltages.

With +10 volts output at 600 mA, the output resistance is 0.07 ohms; ripple is less than 2 mV rms. The output resistance of the slaved negative supply at −10 volts at 600 mA is 0.01 ohms with less than 2 mV ripple. Line voltage regulation is about 1500:1 with either supply. Current limiting in the positive supply is controlled by the series-connected 0.22-ohm resistor; a 1-ohm resistor here will current limit the supply to 200 mA.

The input capacitors specified on the schematic are best buys if you're interested in small size and low price. John Meshna, 21 Allerton Street, Lynn, Mass. 01904, has some big ones for a dollar that are very good, but a new Sprague type 36D 5500 μF only costs $2.25 and is about the size of a 2E26.

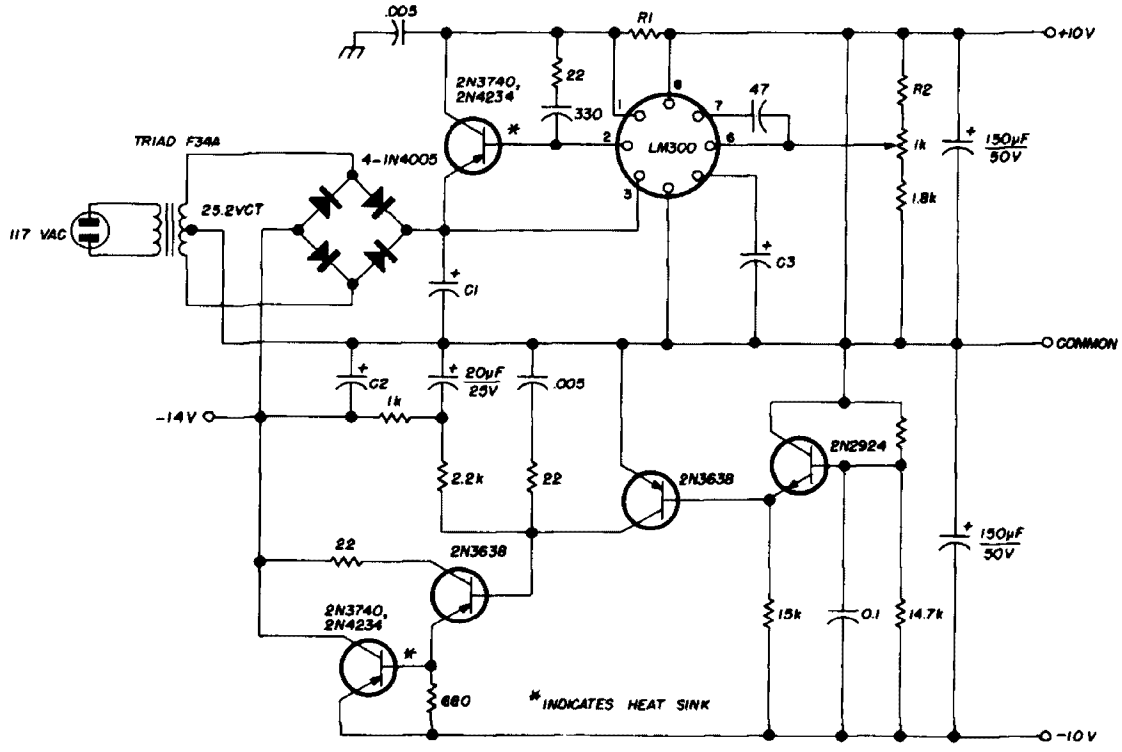
Hank Cross, W1OOP

fig. 2. Slaved power supply provides both positive and negative sources with one voltage-regulator IC. Output voltage depends on value of R2: ±10 V, 600 mA, R2 = 10k; ±12 V, 500 mA, R2 = 13k; ±15 V, 250 mA, R2 = 17k.

C1, C2 value depends on current required. 2000 μF, 25 V for 425 mA; 2700 μF, 25 V for 600 mA; 5500 μF, 25 V for 900 mA (Sprague type 36D or 39D, or Mallory type CG)

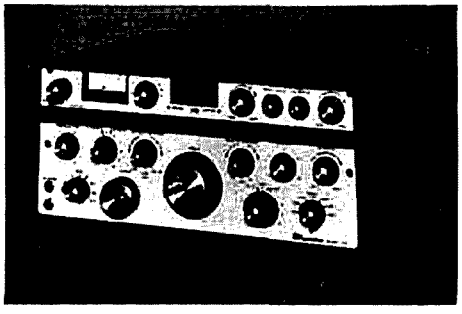
C3 zener noise filter; over 10μF at more than 3 working volts

R1 sets current limit; 0.22 ohm for 600 mA; 1 ohm for 200 mA



new products

signal/one



The new Signal/One CX7 communications package provides a number of operating capabilities heretofore unavailable in an amateur-band transceiver. This "integrated amateur station" package covers seven 1-MHz frequency bands in the 1.8 to 30 MHz range as well as receive-only on 50 to 51 MHz.

The CX7 is completely solid state except for the final power amplifier. The conventional frequency dial is replaced with an integral frequency counter that provides direct frequency readout with 100 Hz accuracy. Two permeability-tuned oscillators provide the 1-MHz tuning range with ample overlap on each band; the tuning rate is 25 kHz per knob revolution. The unit may be operated transceive with either vfo; on split frequencies using both vfo's; transceive on either vfo plus independent receive-only on the other; or transmit-offset transceive where the transmitter frequency tracks the receiver and may be instantly offset up to ± 5 kHz from the receive frequency.

Receiver sensitivity is better than 10 dB signal-plus-noise-to-noise ratio for 1/3 μV input at 29 MHz (2-kHz bandwidth). Excellent selectivity on ssb: 2 kHz at −6 dB and 3 kHz at −60 dB—is a result of two cascaded 8-crystal lattice filters. Ultimate attenuation and image rejection are greater than 80 dB; i-f rejection is in excess of 60 dB and internal spurious are less than 1 μV equivalent.

The hang-type agc system results in less than 6 dB audio output change with signal levels from 1 μV to 100 mV. The hang time is selectable, 100 milliseconds or 1 second. The receiver exhibits good front-end dynamic range; a 10 μV desired signal is modulated less than 10% by an unwanted 10,000 μV signal 10 kHz away (modulated 30% at 400 Hz), or by a similar 100,000 μV signal removed 5%.

Passband tuning is accomplished by varying the actual i-f frequency up to 2 kHz above and below nominal; tracking error (received signal shift) is unmeasurable. The receiver also features dual receive—two receive channels usable separately or simultaneously with a continuously variable relative rf gain control. Either channel may be used for transceiving and the other for receive-only.

The transmitter section of the CX7 is rated at 300 watts PEP input with adjustable rf speech clipping for exceptionally good ssb performance; the transmitter will also operate on CW and fsk. The power amplifier stage is broadbanded for the amateur bands from 1.8 to 29.7 MHz so no tuning is required if the load swr does not exceed 1.5:1. For non-amateur frequencies or excessive swr's, an internal matching network may be switched into the circuit.

Other features of the new CX7 are a built-in noise blanker with adjustable threshold, built-in electronic keyer for 5 to 50 wpm, built-in power supply for 115/230 Vac, push-to-talk or fast-attack vox, instantaneous break-in CW, fast key-up receiver recovery independent of agc hang time, and metering of rf clipping, drive level, plate and screen current and forward and reflected power. Optional accessories include CW and fsk filters, mobile mount and dc adapter, miniature mobile control head, station console and vhf adapter.

Price for the standard CX7 is $1495. For more information, write to the Marketing Manager, Signal/One Division, 2200 Anvil Street North, St. Petersburg, Florida 33710.

75 cents

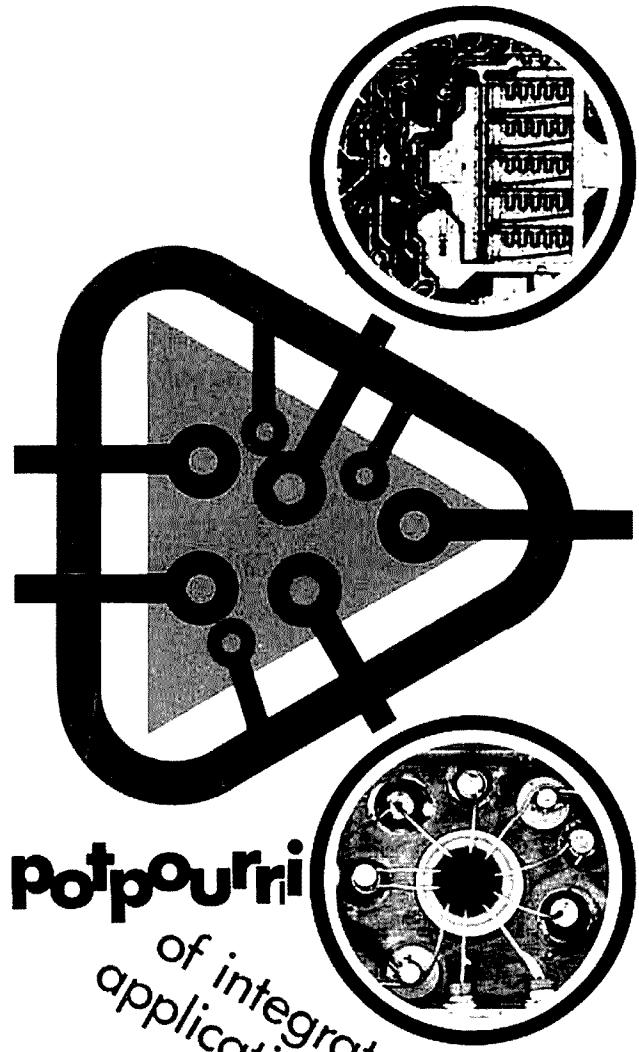
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focus
on
communications
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ham radio

magazine

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staff

editor
James R. Fisk, W1DTY

roving editor
Forest H. Belt

vhf editor
Nicholas D. Skeer, K1PSR

associate editors
A. Norman Into, Jr., W1CCZ
Alfred Wilson, W6NIF
James A. Harvey, WA6IAK

art director
Jean Frey

publisher
T. H. Tenney, Jr. W1NLB

offices

Greenville, New Hampshire 03048
Telephone: 603-878-1441

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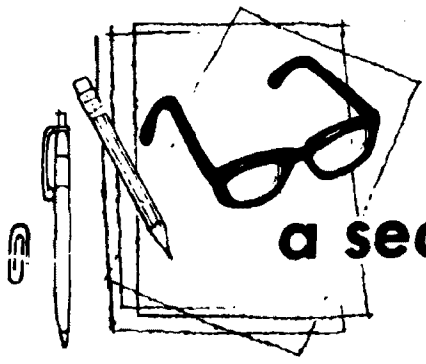
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a second look

by jim fisk

Although the serious DX'ers make up a small percentage of all the licensed amateurs, they are probably more active and make more noise than all the rest of us put together. You may be able to work the world with an S-40A receiver and a DX-20, but you'll never crack a Gus pileup with that gear, even when conditions are right. The big-gun DX'er takes country chasing very seriously and has the QSL's and DXCC to prove it; but what do you do when you've worked them all? Hang up your mike and key and go back to 75-meter phone? Not by a long shot, Charlie! You generate a new award and send everybody back to the starting grid.

Amateur radio covers the world, but there is only one international award, "Worked All Continents," and that was created in 1925 by the International Amateur Radio Union. The number of amateurs who could put "WAC" on their QSL card in the 1920's was pretty slim—WAC was real attainment. Today, if conditions are right, a well-equipped station can work all continents simultaneously in a single roundtable.

In 1966 a group of European amateurs—lead by Gérard de Buren, HB9AW/WA6QAU—got together to set up a new international award that would require activity on all bands. Gérard is station manager of 4U1ITU, the International Amateur Radio Club station in Geneva. Gérard thought it was time to put together an award on an international basis with emphasis on multiband operation; he talked up the idea and was encouraged by interested amateurs all over the world.

He prepared the first draft of rules and presented them to a small working group in London in late 1968. From there on it was up to the 27 board members of the International DX Organization. The rules for the new award, the International Call Areas Award (ICAA), have two very important points: ITU regulations will be followed to the letter and no credit is given for normally uninhabited rocks or reefs. This means that a contact with a station in San Marino using an M1 call will not count for ICAA because it is in the call-sign block allocated to Great Britain. AC4, PX1 and 7G1 fall in the same category. Examples of uninhabited islands that are not on the ICAA list are Bajo Nuevo, Malpelo, Clipperton and Heard Island.

The current ICAA list has a total of 444 call areas: the large number is generated by certain countries that are expanded because of a large amateur population or wide-spaced geographical area. The United States, for example, is broken into the 48 continental states for the purpose of the award. Australia, Canada, Germany, and Japan, as well as several others are "expanded" in a similar way.

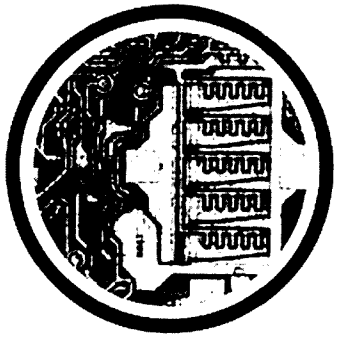
The ICAA is issued in four classes: class I for 400 call areas and 1500 points, class II for 300 call areas and 1300 points, class III for 200 call areas and 900 points, and Class IV for 100 call areas and 500 points. Intercontinental contacts on 1.8 MHz count 10 points (5 points if within the same continent). Contacts on 3.5 MHz count 3 points, contacts on 7 MHz count 2 points and contacts on ten, fifteen and twenty meters count one point per band. All contacts must date after January 1, 1969.

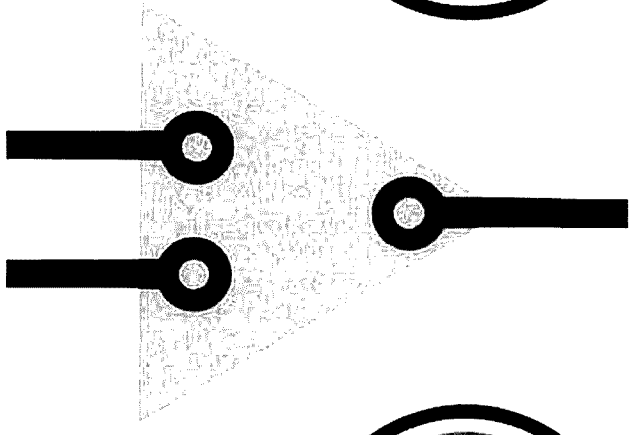
Two official ICAA record books are necessary for each award; one is used for the application and the other for your files. Tentative price for the two record books is $3. In addition, a registration fee of $2 is required from each applicant. If you're interested in chasing a new DX award that is in tune with today's technology, write to the International DX Organization, Post Office Box 543, 1211 Geneva 3, Switzerland.

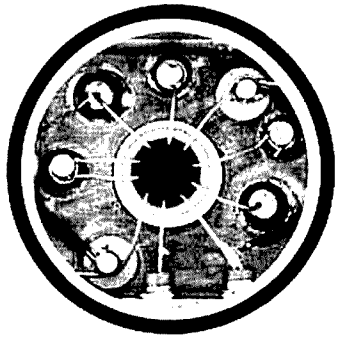
Jim Fisk, W1DTY
Editor

a potpourri of integrated circuit applications

Here is a collection of more than fifty linear integrated-circuit applications for audio, i-f and rf, oscillators, voltage regulators and radio communications





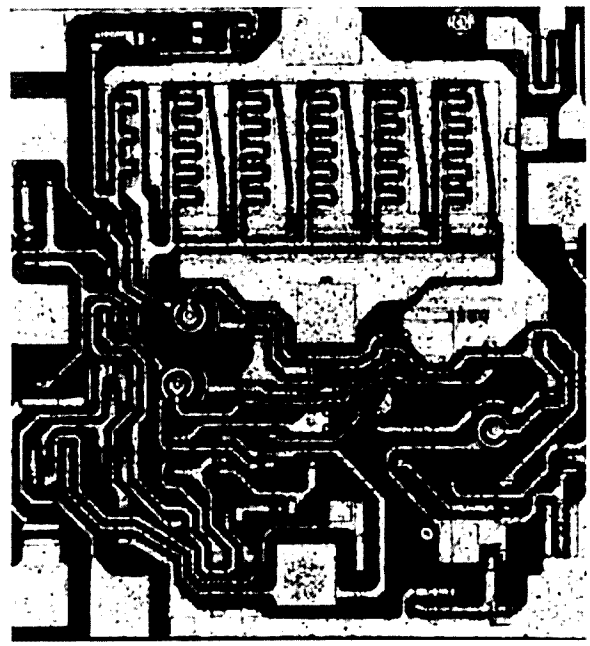


Darrell Thorpe, 3110 N. 83 Street, Scottsdale, Arizona 85251
Jim Fisk, W1DTY, RFD 1, Box 138, Rindge, New Hampshire 03461

Integrated circuits offer many unique possibilities in amateur equipment that are impractical (and often impossible) with conventional solid-state techniques. They simplify many circuits and let you do the same job at less cost—although the innards of one IC may be many times more complex than anything you would be willing to build on your bench.

This article brings together a variety of circuits to give you an idea of what can be done with today's linear IC's. The circuits were collected from a number of different sources including manufacturer's literature, engineering magazines, correspondence with other amateurs and our own basement experiments. The intention is to provide a quick-reference source to integrated circuit applications that are especially suitable for radio communications. With all the integrated circuits that are currently on the market, it is impossible to cover every type; however, we have tried to include a variety of low-cost readily available devices.

If you have never used integrated circuits before, there are several things to keep in mind. First of all, you can't use the same casual haywire breadboard layouts that work ok with transistors and vacuum tubes; if you do, you'll have nothing but trouble. Remember that the IC has a great deal of gain in a very small package. Also, the transistors used in an IC are typically rated to 400 MHz even if the IC is being used as an audio power amplifier. Keep all the leads short, use good rf construction practices in all IC projects and use the bypass capacitors and decoupling resistors recommended by the manufacturer.



Although it is less than 1/8" square, this integrated-circuit die contains 22 transistors (including one power type), 10 diodes and 17 resistors.

Use particular care when making input, output and power supply connections to the IC. Always think in terms of low-resistance and low-inductance grounds and power supply leads. Bypass and frequency-rolloff capacitors should be connected directly at the device socket if possible. Make sure all power supply leads are properly bypassed (at the device), keep input and output leads short and shielded if necessary, and use one common tie point for all grounds. If you follow these simple precautions, you should have a minimum of trouble with integrated circuits "taking off" at some unwanted frequency.

Sockets aren't absolutely necessary, but they are recommended. If you don't use them in your first integrated circuit project, you probably will on later ones! Several manufacturers make sockets, including Cinch-Jones, Elco and Augat; about the most inexpensive sockets are marketed by Motorola with the HEP line; they have sockets for 8- and 10-lead TO-5 cans plus sockets for the 14-lead dual-inline plastic package.

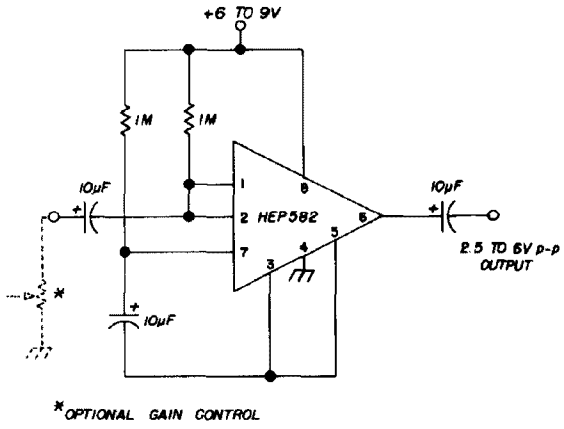


fig. 1. This high-gain audio preamplifier uses a low-cost digital IC—an RTL dual buffer.

audio preamplifiers

One of the simplest audio amplifiers that can be built with an integrated circuit is shown in **fig. 1**. This circuit uses both sections of a digital IC—a dual RTL buffer—to provide gains in the range of 5000 to 10,000 with output voltage swings between 2.5 and 6 volts. The HEP582 dual buffer has the collector load resistors built into the circuit so it's only necessary to add base bias and coupling capacitors to build an audio preamp. This IC really simplifies building a two-stage audio amplifier and is useful right after the detector in a receiver, as a mike preamp or as an oscilloscope preamp to increase scope sensitivity. The HEP582 can also be used as a broadband rf amplifier up to several MHz.

The high-gain low-level preamplifier shown in **fig. 2** uses an HEP592 dual preamplifier that is designed for stereo preamps. This IC is actually a dual operational amplifier, so the operational characteristics of any circuit that it is used in are a function of the external feedback components. The frequency response of the HEP592 can be tailored to your specific application by careful selection of the feedback resistors and capacitors shown in **fig. 2.** The bandwidth parameters of the complete preamp are defined as shown in **fig. 3**; the three "corner" frequencies are set by the external components as indicated by the three formulas.

fig. 2. High-gain preamplifier using one-half of an HEP 592; frequency response is set by the external feedback components as shown in fig. 3.

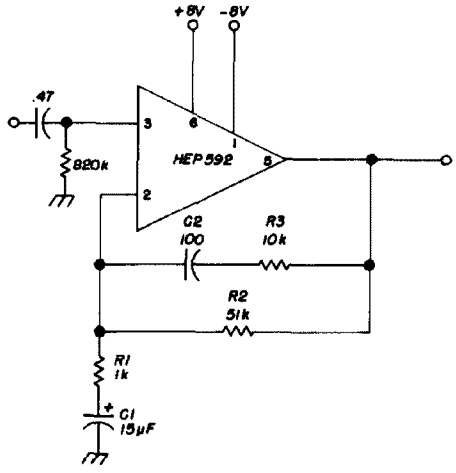
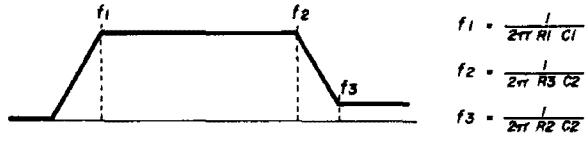


fig. 3. Frequency response and voltage gain of operational amplifiers are set by the external feedback components as shown here.



With the feedback components shown in **fig. 2**, the response is essentially flat from 10 Hz to 100 kHz. Voltage gain at 1 kHz is approximately 40 dB and the maximum

fig. 4. High gain audio preamplifier using a PA230. Frequency response characteristics are determined by the feedback components shown in fig. 3. Table 1 shows typical values and performance characteristics.

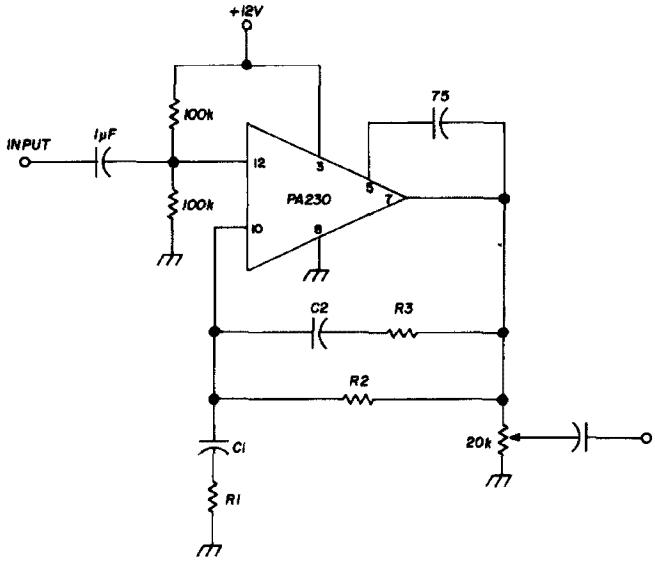
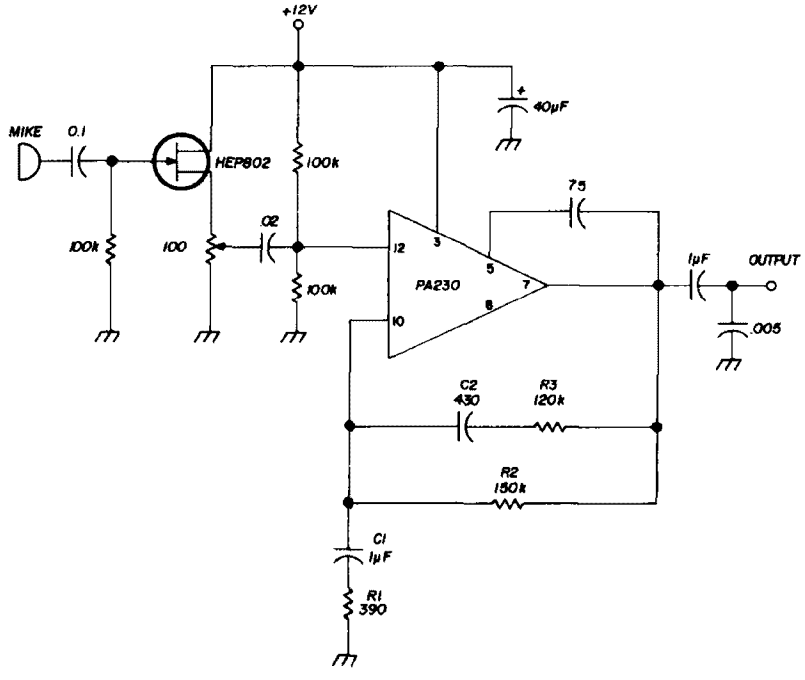


table 1. Frequency response and voltage gain characteristics of the amplifiers shown in figs. 4 and 5.

Bandwidth	Voltage Gain	C1	C2	R1	R2	R3
28 Hz—270 kHz	27 dB	10 μF	10 pF	470	10k	10k
28 Hz—230 kHz	36 dB	10 μF	10 pF	470	27k	10k
28 Hz—80 kHz	38 dB	10 μF	51 pF	470	27k	10k
28 Hz—25 kHz	41 dB	10 μF	100 pF	470	56k	10k
310 Hz—2700 Hz	40 dB	1 μF	430 pF	390	150k	120k

The high-gain audio preamplifier shown in **fig. 4** uses a General Electric integrated circuit, the PA230. This IC features expanded operating temperature range, output short-circuit protection, high voltage gain and low noise. The frequency characteristics of the preamp circuit may be set using the criteria shown in **fig. 3**. **Table 1** shows the frequency response and voltage gain characteristics with various combinations of feedback resistors and capacitors. The circuit in **fig. 5** uses a PA230 and an inexpensive fet to provide a high input impedance high-gain microphone preamp for communications work; the frequency response of this circuit is approximately 300 to 2700 Hz.

fig. 5. High input impedance microphone preamplifier with frequency response suitable for ssb.



output voltage swing is 5 volts rms. Since only one-half the HEP592 is used in **fig. 2**. the other half (which is identical) may be used for additional amplification or for another amplifier channel. Channel separation is typically 58 dB.

audio power amplifiers

If you need a miniature half-watt audio power stage, the circuit in **fig. 6** should be of interest. This circuit uses a single IC—the RCA CA3020 or CA3020A—to provide 545 mW output with 45 mV input. The output from a solid-state product detector or a-m detector usually provides enough drive for full output. In this circuit the idling current is 22 mA, input resistance is 50k ohms and total harmonic distortion at 135 mW output is 3.3%. For more information on uses of the CA3020, see reference 18.

The 1-watt audio power amplifier shown in **fig. 7** is about the most simple integrated-circuit power amplifier you're apt to see. The PA234 is designed to operate with 9- to 25-volt power supplies into 8-, 16- or 22-ohm loads. The voltage gain of the circuit depends upon the ratio of R2/R1; biasing is set by the ratio R2/R3. In addition to the resistors, three capacitors are required for input and output coupling and high-frequency stabilization.

The ratio of R2/R3 must equal 10 for proper bias with a 22-V supply and to maintain good stabilization, R3 should not be greater than 100k. With the values shown in the circuit diagram, voltage gain is approximately 8 and a 600 mV input signal is required for 1 watt output. Distortion at 1 watt output is 3%; at 0.05 watt output distortion is 0.5%.

When using the PA234, the power supply output filter should be as close to pin 10 as possible with **all** other ground returns run separately to this point because common ground impedances can cause hum and distortion. Since the frequency response of the amplifier extends from 30 Hz to 100 kHz, the 0.068-μF capacitor across the output is required for protection against oscillation.

The Amperex TAA300 integrated-circuit audio power amplifier circuit shown in **fig. 8** can deliver one watt into an 8-ohm load when powered from a 9-volt power supply. In this circuit the total harmonic distortion at 1 watt output is 10%; total harmonic distortion at 0.5 watt is 3% maximum. Input impedance is greater

fig. 7. One-watt audio power stage using a PA234. Heat sink tab must be soldered to heat sink area not less than 2 square inches.

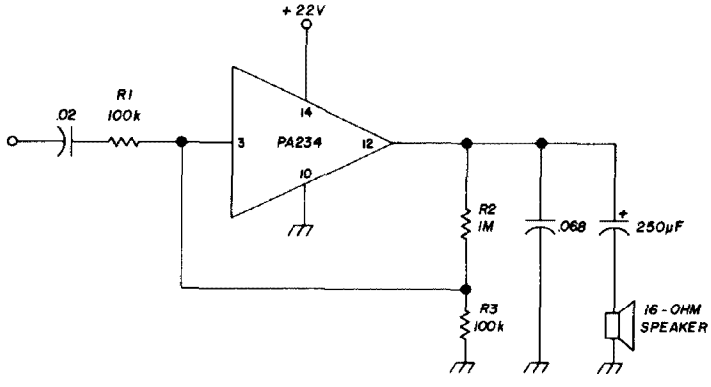
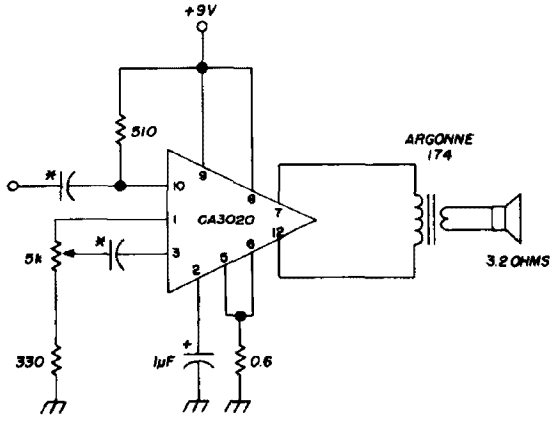


fig. 6. Half-watt audio power stage. For most amateur applications the capacitors marked with an asterisk should be 0.25 μF; for better low frequency response, use 1-μF nonpolarized capacitors.



than 10k ohms, and a 10 mV input signal will provide 0.7 watt output. A finned heat sink must be used when the TAA300 delivers one watt into an 8-ohm load; no heat sink is required if the load impedance is 16 ohms or greater.

Although this circuit is designed for a +9-volt supply, the IC will operate satisfactorily with power supplies down to +4.5 V. The variable resistor, R1, is adjusted for 8 mA total current drain with no

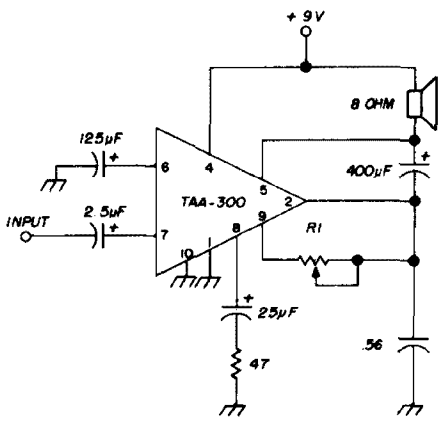


fig. 8. Amperex TAA300 1-watt audio amplifier. R1 is adjusted for 8 mA total current drain with no input signal.

input signal. When the circuit is fed from an ac power source, typical value for this resistor is 330 ohms; when a battery supply is used, it is preferable to use a smaller value because of the resistor's affect on operating current stabilization at low battery voltages.

Do not short or use too low a resistance value for R1 when **setting** the no-signal operating current. Start with a large resistance—approximately 25k—and reduce the value in small steps to arrive at the correct operating point. A 2.2k ohm resistor should be connected in series with R1 initially to protect the circuit in case the adjusting potentiometer is rotated to the zero resistance position.

The audio power amplifier shown in **fig. 9** uses an integrated circuit that was designed to amplify signals to 300 kHz with 1.8 watt delivered into a direct or capacitively coupled load. The HEP593 IC features low harmonic distortion—0.4% typical at 1 watt—plus low output impedance and excellent gain-temperature stability. The voltage gain of this power amplifier stage is determined by the connections to the "gain-option" pins. For a voltage gain of 10, pins 2 and 4 are open and pin 5 is connected to **signal** ground; for a voltage gain of 18, pins 2 and 5 are open and pin 4 is connected to signal ground; for a voltage gain of 36, pin 2 is connected to pin 5 and pin 4 is connected to signal ground.

To avoid vhf instability with this circuit the RC stabilizing network—0.1 μF in series with 10 ohms—must be placed directly from pin 9 to ground with **short** leads. Excessive lead inductance from the positive supply to pin 10 can cause high-frequency instability. The B+ bypass capacitor should be connected directly from pin 10 to ground if possible; if not, the series RC network shown in **fig. 9** should be used

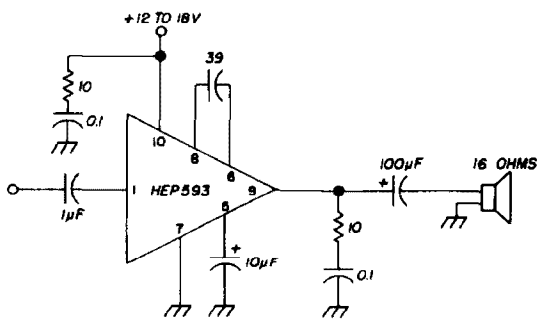


fig. 9. This audio power amplifier will provide up to 1.8 watts output. Voltage gain is determined by connections to pins 2, 4 and 5 as discussed in the text.

directly from pin 10 to ground. In addition, lead lengths from the external components to pins 7, 9 and 10 should be as short as possible to insure good vhf grounding at these points.

Because of the large bandwidth of the HEP593, coupling must be avoided between the input and output leads. This can be accomplished by using short leads which are well isolated, narrow-banding the over-all amplifier by putting a .001

table 2. Effect of feedback resistance (Rf in fig. 10) on sensitivity, distortion and input impedance.

Rf	Sensitivity at 2 W	THD	Input Impedance
0	8 mV	5.2%	20k
1k	22 mV	2.9%	35k
5k	86 mV	1.7%	40k
6.8k	120 mV	1.6%	40k
10k	150 mV	1.5%	40k

capacitor from pin 1 to ground, and by using shielded input cable.

The audio power circuit shown in **fig. 10** can deliver 2 watts into a 16-ohm load. The PA237 may be operated with supply voltages from 9 to 27 volts and is capable of 1 ampere peak output. In the circuit of **fig. 10** the integrated circuit is biased into class AB. The voltage gain of the circuit as shown here is greater than 45; a 120 mV input signal will produce two watts output. The input impedance is 40k ohms, output impedance is 0.85 ohm and noise output is 75 dB below 2 watts.

By setting the 6.8 kilohm feedback resistor (Rf) to other values, the sensitivity, distortion and input impedance will vary as shown in **table 2**. Since distortion decreases when the output level is decreased by increasing the negative ac feedback, it might be a good idea to use this feedback resistor as the volume control. This is particularly desirable for lowering crossover distortion at low levels where it becomes a significant part of over-all amplifier distortion.

In **fig. 11** the PA237 is used in an audio power amplifier that will provide 2 watts into a 16-ohm load with 5 mV drive; this is a voltage gain greater than 1100. With this amount of sensitivity, the amplifier can be driven directly with a microphone or low-level detector. The same circuit can be used with other combinations of voltage and load impedance by changing the values of the bias resistors; **fig. 12** shows the same basic circuit with an 8-ohm load and 12-volt power supply. The voltage gain of this circuit is 350, and 7 mV input drives the amplifier to 0.75 watt output. The input impedance of the circuits shown in **fig. 11** and **12** is approximately 15k.

If you need more than the two-watt

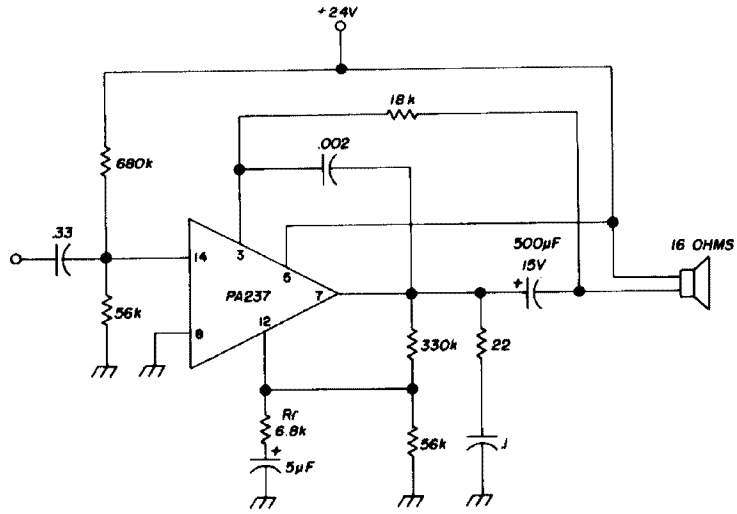


fig. 10. Two-watt audio power amplifier. Circuit sensitivity is determined by feedback resistor (Rf) as shown in table 2.

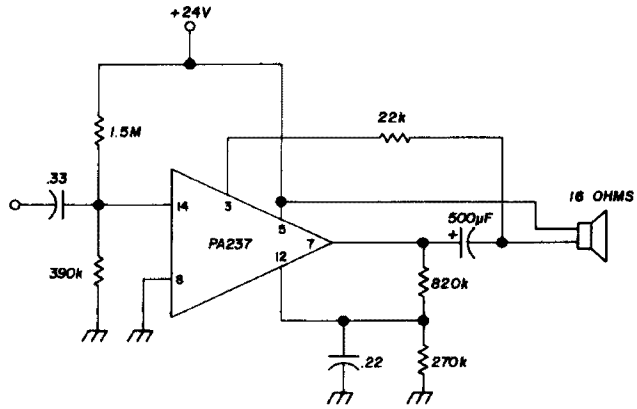


fig. 11. This power amplifier provides 2 watts output with 5 mV drive.

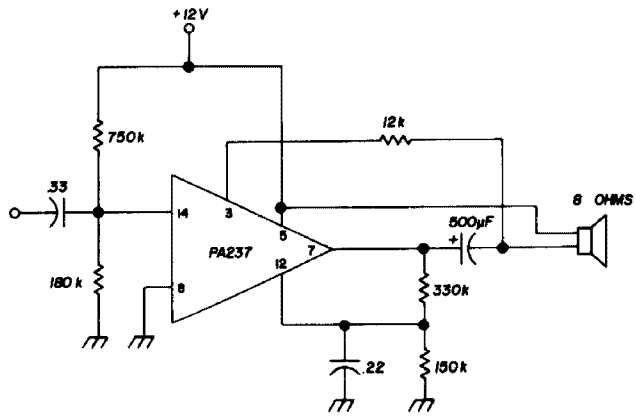


fig. 12. This circuit is the same as fig. 11 except that it is set up for a 12-V supply and 8-ohm speaker; 7 mV input drives it to 0.75 watt output.

capability of one PA237 integrated circuit, two of the devices may be connected in the bridge configuration shown in **fig. 13.** This circuit effectively doubles power output. Since each circuit is the same as **fig. 10** the feedback resistances given in **table 2** may be used to set amplifier input sensitivity and input impedance.

The highest power audio amplifier IC currently available is the new General Electric PA246 with a capability of 5 watts continuous into a 16-ohm load with less than 1% total harmonic distortion. In the circuit shown in **fig. 14,** 180 mV input will provide 5 watts output at 0.7% total harmonic distortion. When the feedback resistor (Rf) is shorted out, sensitivity increases; 12 mV input provides 5 watts output. Frequency response extends from 30 Hz to 100 kHz, and noise output is typically 70 dB below 5 watts.

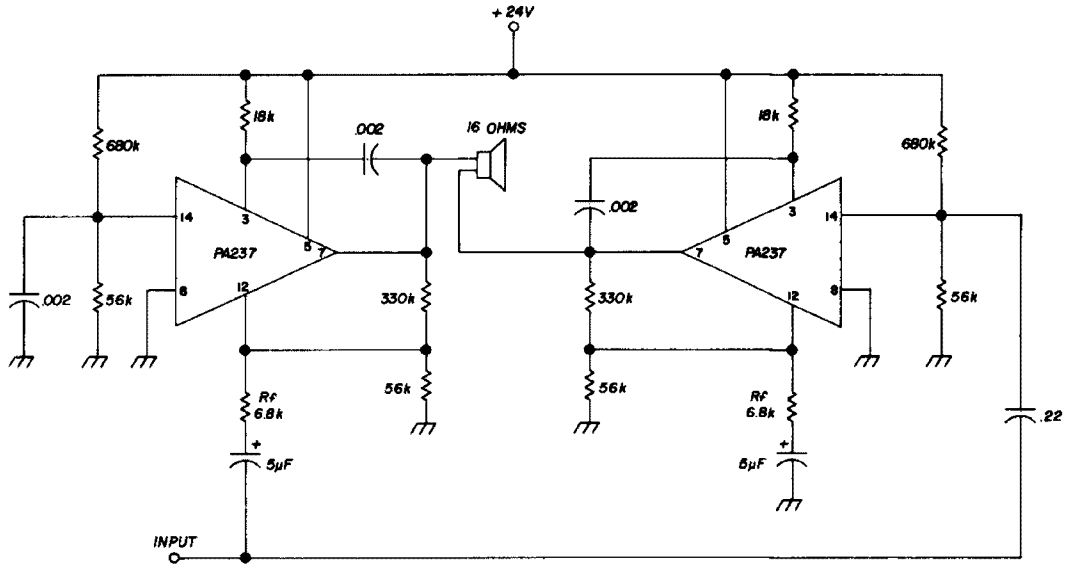


fig. 13. Two PA237's in a bridge double power output to 4 watts.

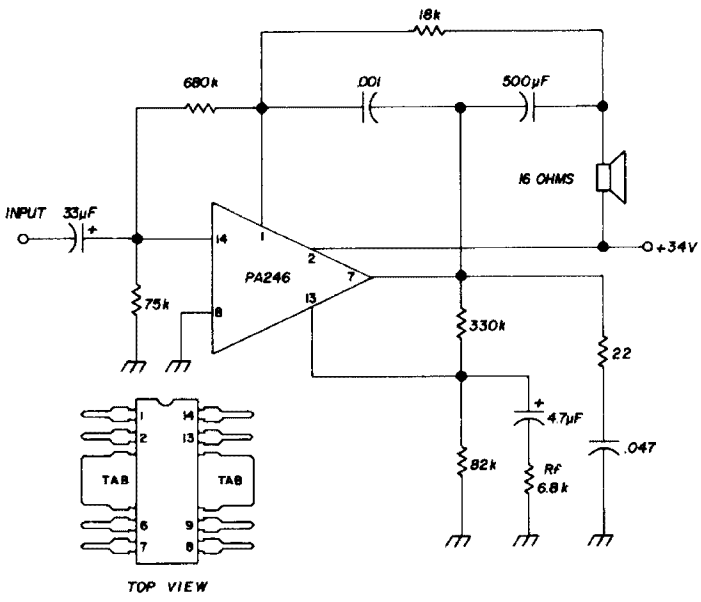


fig. 14. Five-watt audio power amplifier. The speaker should be a permanent-magnet type so resistive load is close to 16 ohms. Heat-sink tab should be connected to pin 8, circuit ground.

The increased power dissipation capability of the PA246 is a result of a higher permissible supply voltage—up to 37 volts —and increased heat dissipation with two large heat-sink tabs. The higher voltage rating allows more power without increased heat-generating current and may eventually lead to audio power IC's with ratings up to 10 watts.

Another way to gain increased power output is to use the approach of **fig. 15.** In this circuit a low-level IC preamp is used to drive a discrete complimentary output stage: a Fairchild μA716C is used as a driver for an output stage that will put 2 watts into a 16-ohm load. The fre-

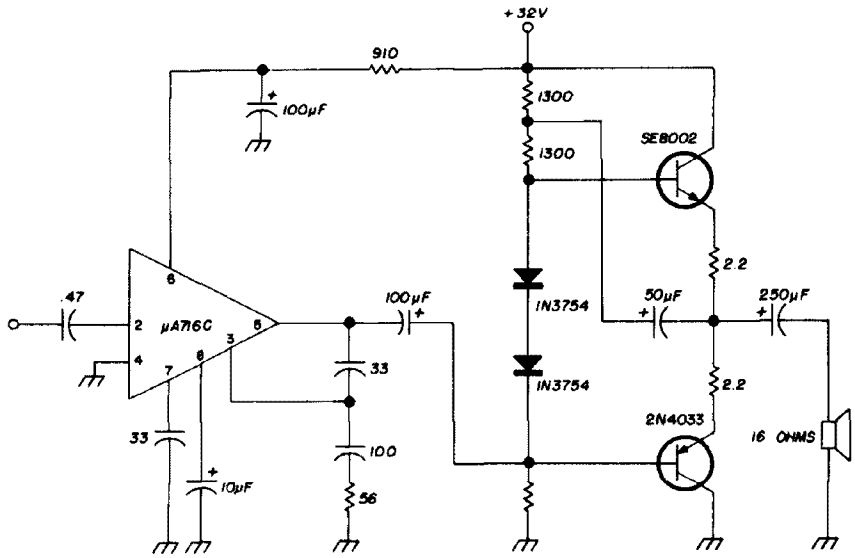


fig. 15. Two watt audio power amplifier with complimentary transistor output stage.

quency response of **fig. 15**—in reference to 1000 Hz—is −1.5 dB at 100 Hz and −4 dB at 20 kHz. Although Fairchild transistors are shown here, RCA devices such as the 40053 npn and 40319 pnp may be substituted. Dc current gain (h_{FE}) should be in the range of 60 to 100 and nearly the same in both transistors to prevent waveform clipping.

Although the highest power audio amplifier IC currently on the market is General Electric's 5-watt PA246, Bendix Semiconductor has recently announced a 15-watt audio amplifier IC that uses thick-film construction. It is not nearly as small as the other audio power IC's mentioned here, 1 x 2 x 5/16 inch, but it is still useful for many amateur applications. The 15-watt amplifier, the BHA0002, has a frequency response of 25 Hz to 20 kHz, power gain of 55 dB, total harmonic distortion less than 1%, supply voltage range of 14 to 40 volts and sensitivity of 350 mV for 15 watts output.

audio mixer

The four-input audio mixer shown in **fig. 16** may be used for mixing low-level audio signals up to 50 kHz. With a 6-volt power supply the input will handle up to 300 mV peak-to-peak; maximum output is 3 volts p-p. The 5000-ohm potentiometer should be adjusted for minimum distortion.

audio limiter

The low-distortion audio limiter shown in **fig. 17** begins to limit at 0.4 millivolt, is not affected by input signals up to 6 volts p-p and operates linearly below the clipping point without oscillation or other instability. A good limiter should limit on millivolt signals but not distort at higher levels; it should be able to reproduce a square wave with very little even-harmonic content while providing clean sine waves below the limiting threshold. By using a high-gain operational amplifier with an external clipper and feedback circuit, second harmonic generation is less than 0.3% over a dynamic range of 54 dB and less than 2% over a 78 dB range.

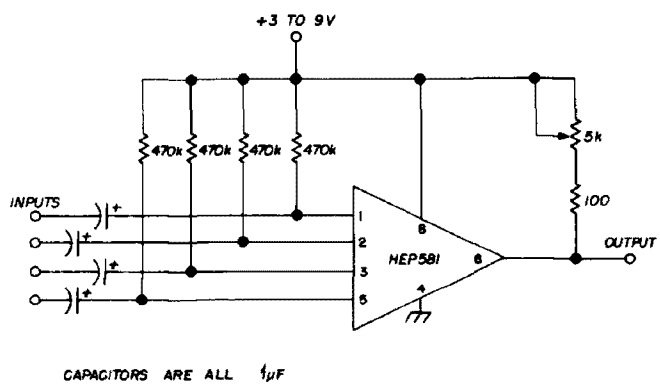
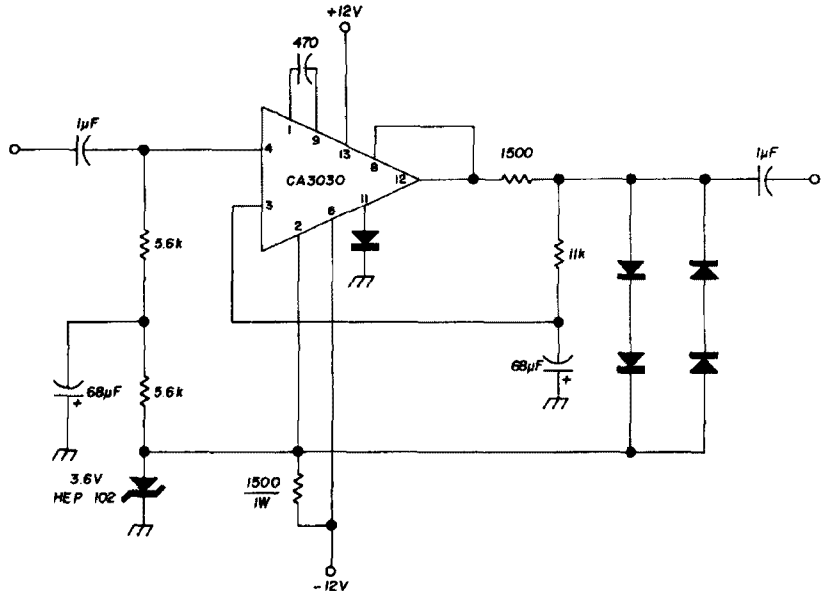


fig. 16. Four-input audio mixer using a digital 4-input gate. The 5k pot is adjusted for minimum output distortion.

fig. 17. Low-distortion high dynamic range audio limiter.



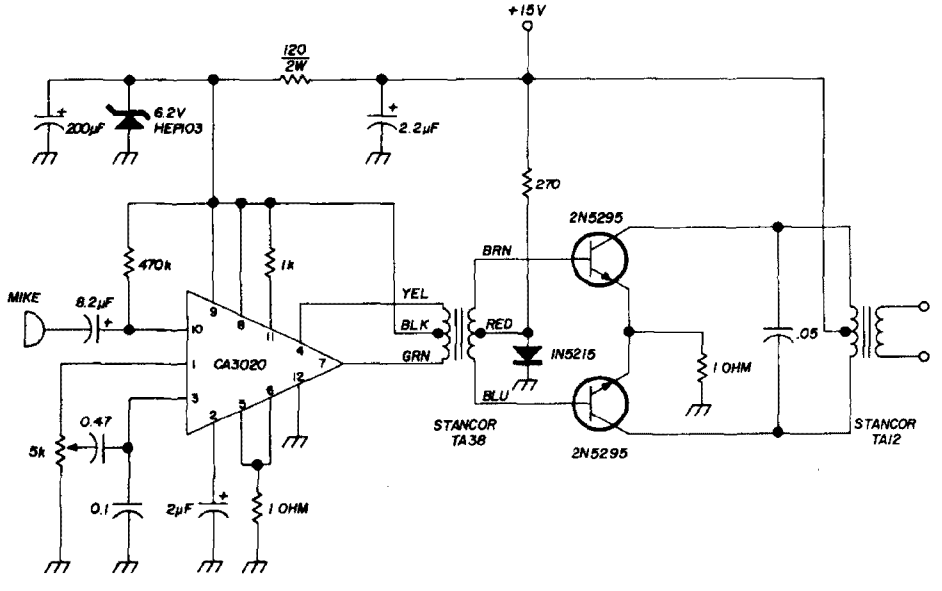
modulator

The modulator circuit shown in **fig. 18** will produce up to 5 watts of undistorted audio. The modulation transformer—a Stancor TA12—is an audio output transformer designed to work into an 8-ohm load so it is most suitable for a transistor rf power stage. The 0.05-μF capacitor between the collectors of the push-pull 2N5295 power stage should be adjusted for minimum distortion and best speech quality with your particular microphone. This circuit is designed for a high output crystal or ceramic microphone; for low-level dynamic microphones, another stage of amplification may be required at the input. If another modulation transformer is used, keep in mind that the 2N5295's should work into approximately a 20-ohm load.

i-f and rf amplifiers

Integrated circuits that are suitable for rf and i-f applications are not actually that much different from the audio-frequency variety. These IC's usually have two or more transistors on a single semi-

fig. 18. Five-watt modulator for transistor rf power amplifiers.



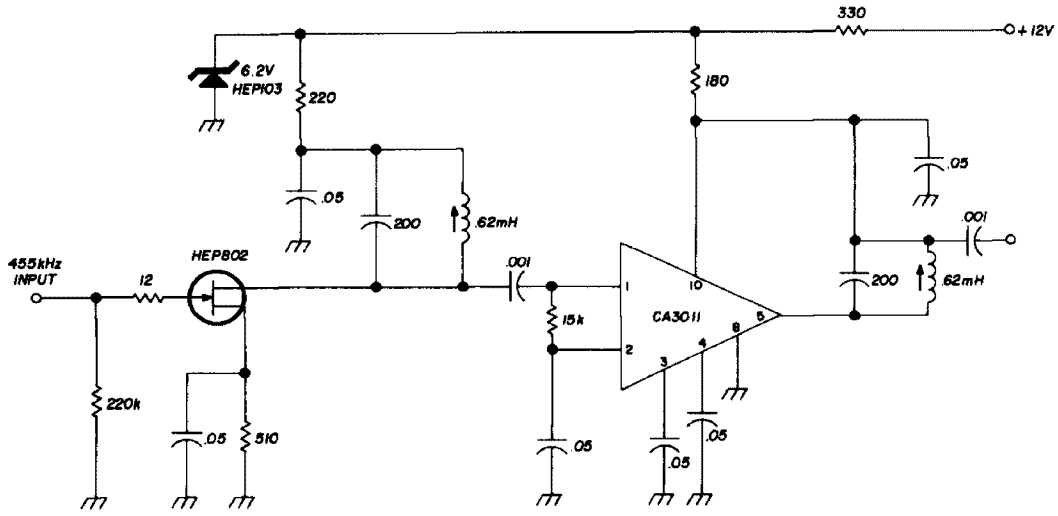


fig. 19. This 455-kHz i-f amplifier uses an fet to provide a high impedance input.

conductor chip along with the necessary biasing network. In some cases several stages of amplification are built on a single chip to provide very high gains in a tiny package. It wasn't too long ago that IC's suitable for high-frequency work were priced beyond the average amateur, but today there are a number of inexpensive devices available that do an excellent job up to 60 MHz or so.

There are both advantages and disadvantages to using IC's in rf and i-f amplifiers; and the direction in which the scale tips is generally related to the frequency of the application. On the plus side, IC's provide high gain figures, are temperature compensated and are easy to use with agc; in addition, they usually do not require external biasing networks. However, current rf-rated integrated circuits that are on the market, are frequency limited—most go to pot if you try to use them much above 100 MHz. There are IC's available that work well into the uhf range, but most of these are engineering prototypes that haven't reached production. As the manufacturing processes are perfected and designs are improved, uhf-rated integrated circuits should become commonplace.

The 455-kHz i-f amplifier shown in **fig. 19** costs less to build than an equivalent circuit built with individual components. The tuned coils are from small i-f transformers used in minature transistor broadcast radios; the low-impedance output windings are not used. The integrated circuits have only a moderately high input impedance so a field-effect transistor is used to keep loading on the input circuitry to a minimum. To modify this circuit to a limiting amplifier for fm use, connect an identical CA3011 amplifier stage to the stage shown here.

Another 455-kHz amplifier circuit is shown in **fig. 20**. Each of the stages represents one-half of a single IC package, so this high-gain amplifier is extremely compact. Over-all gain of this circuit is 67 dB, the 3-dB bandwidth is 3 kHz and the input impedance is approximately 30 kilohms. With the output transformer shown here, the output impedance is approximately 800 ohms; for a 5-kilohm output impedance, use a Miller 2063 transformer. For maximum gain, the agc input should be grounded.

The 10.7-MHz amplifier shown in **fig. 21** provides from 60 to 65 dB gain. The main disadvantage of this circuit is the low number of tuned circuits—it may be difficult to get the amount of selectivity you need. However, this may be solved by using a crystal filter or cascaded tuned circuits. Because of the high gain, this circuit may be unstable unless you are very care-

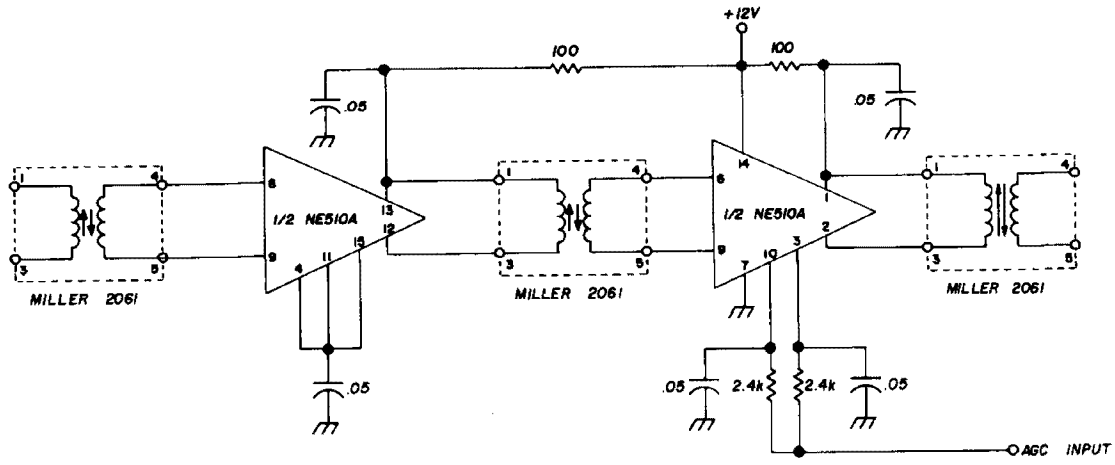
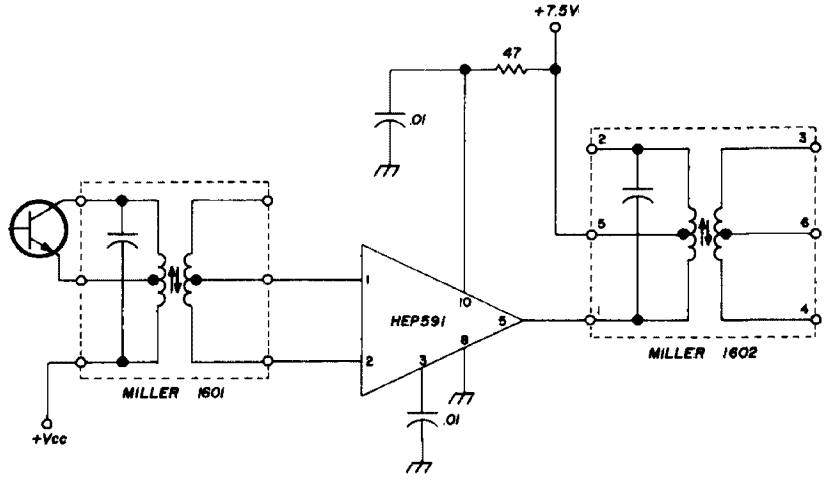


fig. 20. Two-stage 455-kHz i-f amplifier uses one integrated circuit.

fig. 21. This 10.7 MHz amplifier provides gain in the neighborhood of 60 dB. The RCA CA3012 will also work in this circuit.



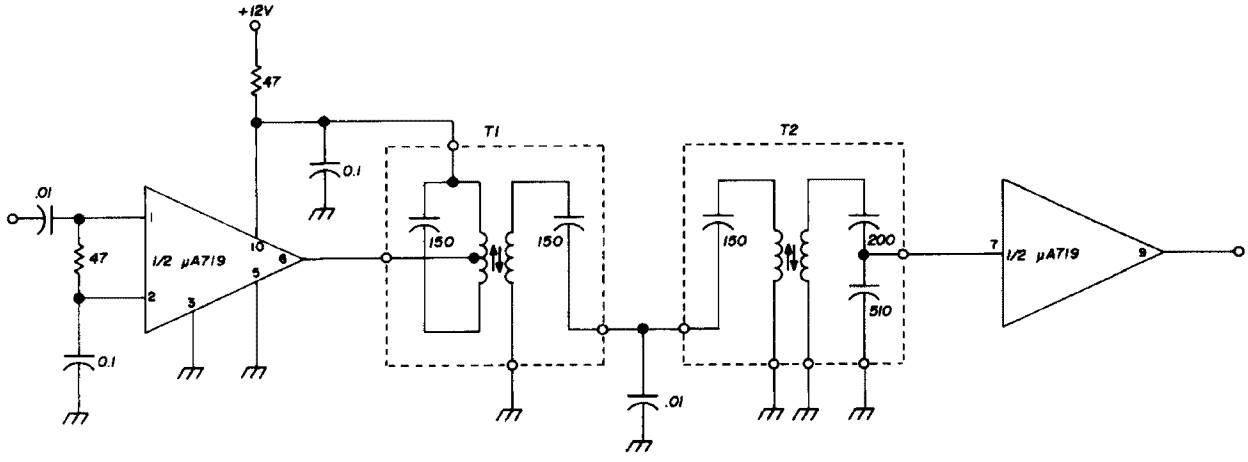


fig. 22. This 10.7-MHz i-f amplifier uses cascaded tuned circuits for improved selectivity.

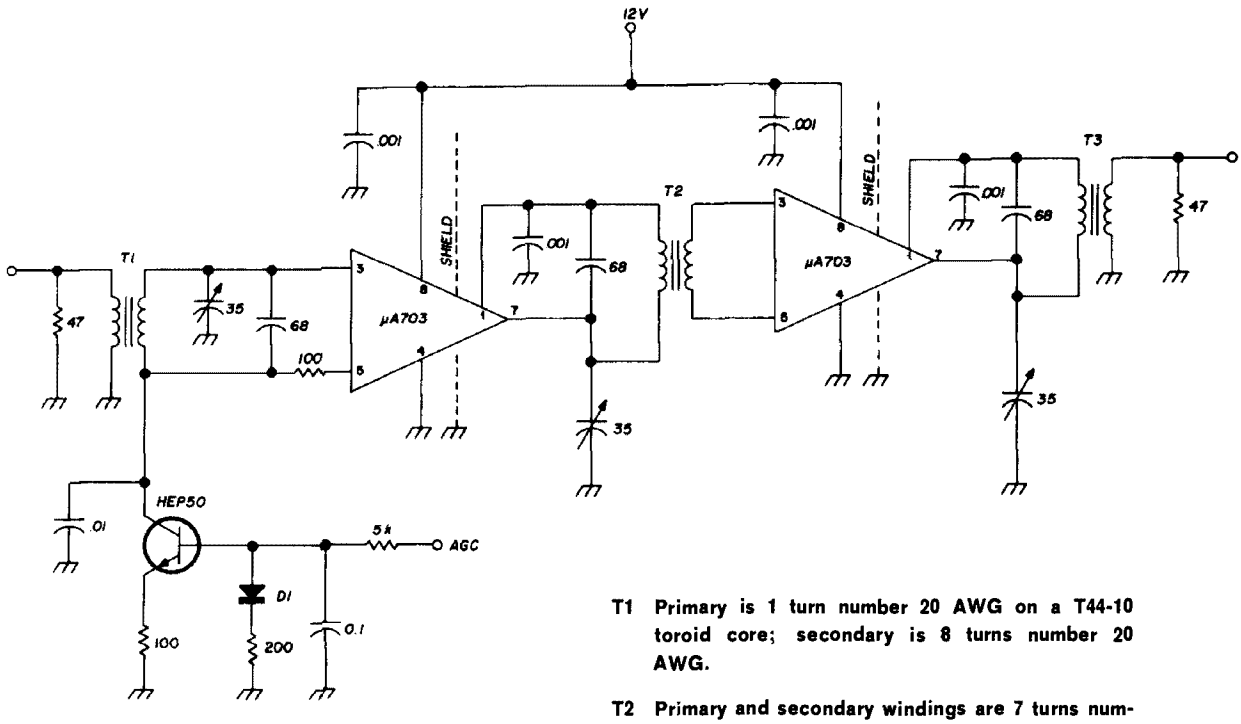


fig. 24. High-gain i-f strip has agc input. Diode D1 may be any silicon diode.

T1 Primary is 1 turn number 20 AWG on a T44-10 toroid core; secondary is 8 turns number 20 AWG.

T2 Primary and secondary windings are 7 turns number 22 AWG bifilar wound on a T44-10 toroid.

T3 Primary is 8 turns number 20 AWG on a T44-10 toroid core; secondary is 1 turn number 20 AWG.

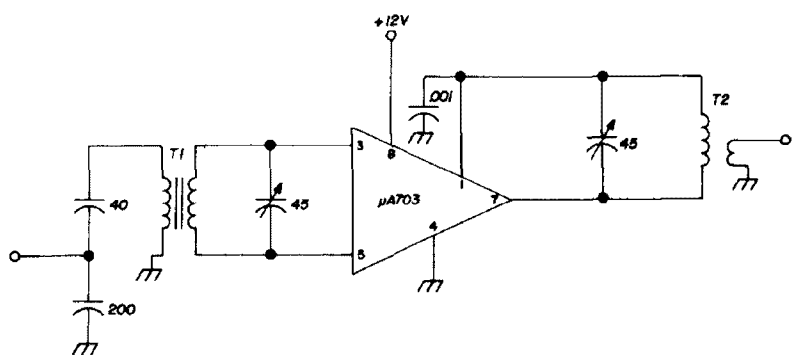
ful when laying it out. Make all the leads as short as possible. Note that the bypass capacitors are somewhat smaller than those usually recommended for this frequency; this reduces gain and helps overall stability.

The 10.7 MHz i-f amplifier circuit shown in **fig. 22** uses cascaded tuned circuits to provide selectivity. The output from the four-pole filter feeds into the second half of the Fairchild μA719. The insertion loss of the filter is such that the second stage of amplification is not driven into limiting.

The primary of T1 consists of 16 turns number 34 enamelled, close wound on a 1/4-inch form, centertapped; secondary is 16 turns number 34 enamelled, close wound on the same form with 0.225 inch between windings. The primary of transformer T2 consists of 16 turns number 34 enamelled close wound on a 1/4-inch slug-tuned form; secondary is 16 turns number 34 enamelled close wound on the same form, tapped 4 turns from cold end; spacing between windings is 0.225 inch. Transformer cans are Miller S-34.

The simple 30-MHz amplifier stage shown in **fig. 23** provides approximately 30

fig. 23. Simple high-gain 30-MHz i-f strip provides 30 dB gain.



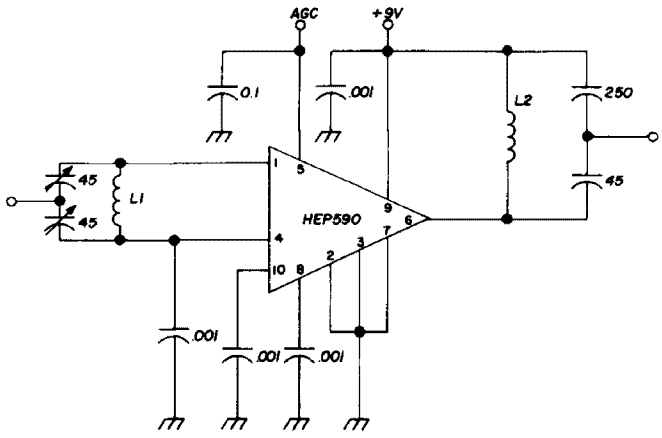


fig. 25. 50-MHz amplifier provides 30 dB gain.

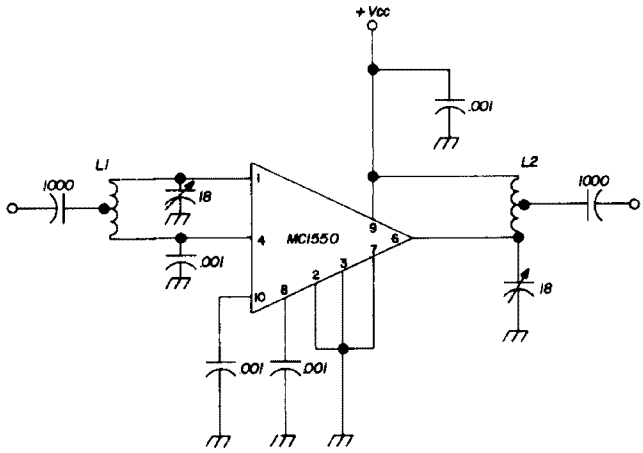


fig. 26. Two meter rf stage exhibits 13 dB gain and 6.8 dB noise figure. The HEP590 will also work in this circuit.

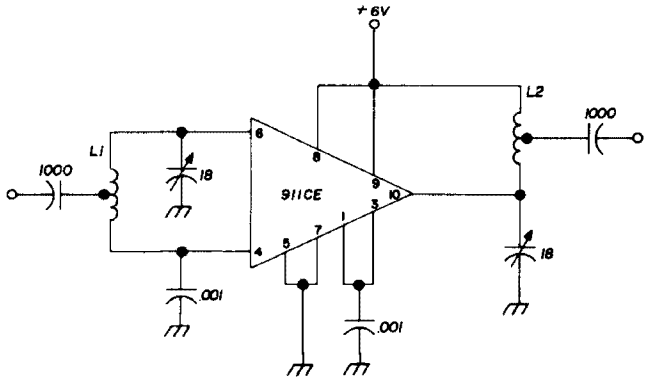


fig. 27. This 144-MHz amplifier will provide up to 17.7 dB gain with a 4 dB noise figure.

dB gain. The noise figure of this stage is 6 dB and the bandwidth is 1 MHz. The primary and secondary of transformer T1 are each 10 turns number 22 enamelled bifilar wound on a T44-10* toroid core. The primary of transformer T2 is 12 turns number 22 enamelled on a T44-10 core; secondary is one turn to work into a 50-ohm load. This stage can also be used as a limiting amplifier for fm work—simply change the turns ratio of the output transformer by making the secondary of T2 two turns (to work into a 50-ohm load). This increases the load impedance the integrated circuit works into and turns the stage into a limiting amplifier.

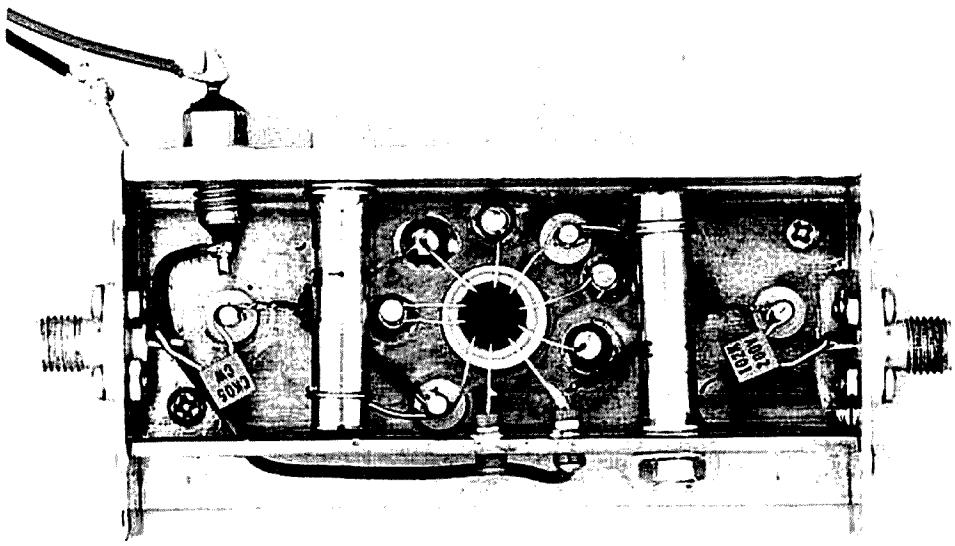
The 30-MHz i-f strip shown in **fig. 24** is an extension of the circuit of **fig. 23** with provision for agc. The agc range is approximately 40 dB with agc voltage variation from zero to 4 volts. Gain of the i-f strip is approximately 60 dB, bandwidth is 900 kHz and noise figure is about 6 dB. For an additional 30 dB gain, another μA703 stage may be added between transformer T2 and the second μA703; the interstage transformer for the additional stage would be the same as T2.

vhf amplifiers

For high-frequency applications of modern integrated circuits, you might consider the vhf amplifier circuits shown in **fig. 25, 26** and **27**. The 50-MHz rf stage in **fig. 25** uses a readily available IC to provide up to 30 dB gain with 18 dB agc range. When the agc voltage is zero, gain is 30 dB; when the agc voltage is 4 volts, gain is 12 dB. Both input and output coils L1 and L2 are 9 turns number 20 AWG on a T44-10 toroid core.

The two-meter amplifier in **fig. 26** exhibits about 24 dB gain with a 25 volt supply; gain drops to 13 dB when the supply voltage is lowered to 6 volts. The best noise figure can be obtained with the 6-volt power supply—6.8 dB. Input coil L1 is

* Toroid cores listed in this article are available from Circuit Specialists Company, Post Office Box 3047, Scottsdale, Arizona 85251; price, two cores $1, postpaid in the U.S.A.



Layout for the amplifiers shown in figs. 26 and 27.

13 turns number 28 AWG on a T20-10 toroid, tapped 6 turns from the cold end; output coil L2 is 13 turns number 28 AWG on a T20-10 toroid, tapped 9 turns from the cold end.

The two-meter amplifier shown in **fig. 27** provides somewhat more gain than **fig. 26**—19.8 dB with a 12-volt supply and 17.7 dB with a 6-volt supply. Best noise figure is again coincident with the 6-volt supply and is about 4 dB. Input coil L1 is 12 turns number 28 on a T20-10 toroid, tapped 5 turns from the cold end; output coil L2 is 11 turns number 28 AWG, tapped 6 turns from the cold end.

The novel integrated-circuit amplifier shown in **fig. 28** may be used as an rf or i-f amplifier at frequencies up to about 200 MHz with 60-dB gain; the IC is a Sylvania SA-20. The circuit is unusual because of the placement of the tuned circuit—as a frequency selective feedback loop between two of the transistors within the integrated circuit. L1 and C2 should resonate at the desired operating frequency, and C1, which is a dc blocking capacitor, is made large enough to prevent series resonance with L1 (at least in the rf range). A notch amplifier results if L1 and C1 are resonated and C2 is removed. Replacing the LC circuit with piezoelectric crystals or ceramic filters results in sharper selectivity.

mixers

Most of the integrated circuits that are designed for rf and i-f applications may also be used as frequency mixers. Most amateurs are not aware of one of the advantages of using a differential amplifier as a frequency mixer—the local-oscillator frequency may be one-half the required injection frequency. At the higher frequencies this is particularly helpful since the ease of building stable oscillators declines rapidly as frequency is increased above 60 MHz or so.

A mixer circuit for the μA703 is shown in **fig. 29.** L1-C1 are tuned to the signal

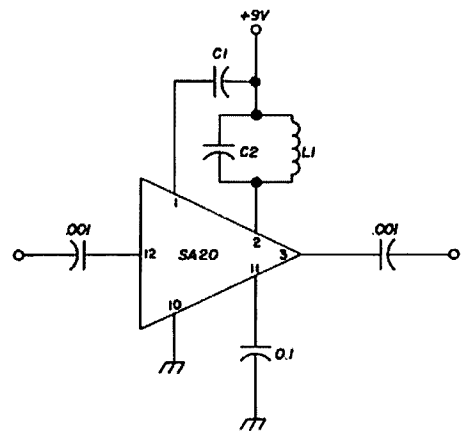


fig. 28. Multipurpose rf/i-f amplifier has good gain and bandpass characteristics up to 100 MHz.

frequency, L2-C2 are tuned to the local oscillator frequency and L3-C3 are tuned to the intermediate frequency. Other rf-rated integrated circuits can be used in this manner as long as you use the differential inputs to the IC. This mixer circuit can theoretically be operated at any **even** harmonic of the local oscillator, although efficiency drops drastically at more than the fourth harmonic.

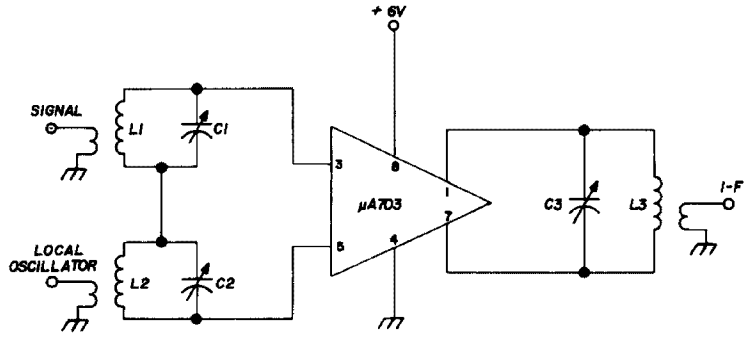


fig. 29. With this harmonic mixer circuit, the local oscillator can operate at one-half the required injection frequency.

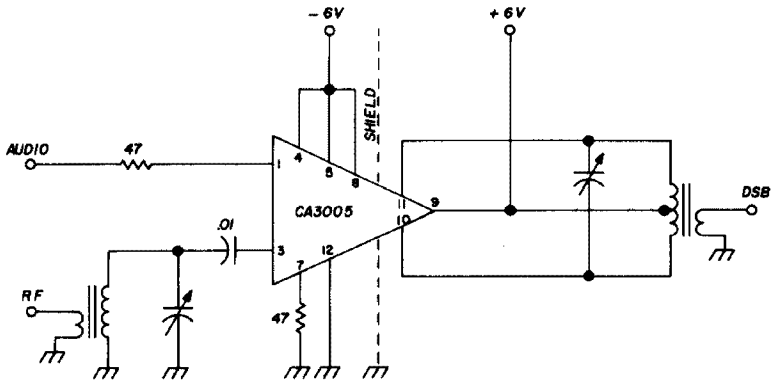


fig. 30. Balanced modulator circuit provides up to 25 dB carrier suppression.

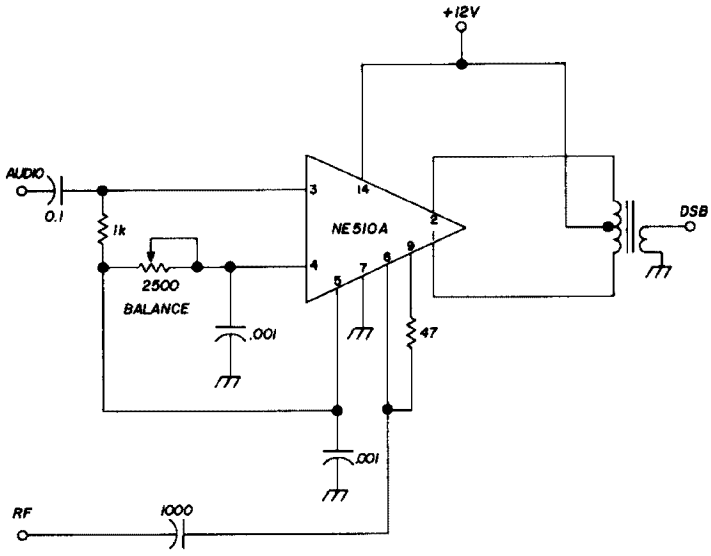


fig. 31. Balanced modulator circuit using a Signetics NE510A.

The balanced mixer (or balanced modulator) shown in **fig. 30** is useful for generating a double-sideband signal. Carrier suppression with this circuit is a function of circuit symmetry and the modulation-to-carrier drive ratio. If external component symmetry is watched carefully, carrier suppression is about 25 dB with 10 mV audio and 32 mV carrier signals. For best results, the output transformer should be bifilar wound.

Another balanced modulator circuit is shown in **fig. 31**. This circuit uses a Signetics NE510A. When the 2500-ohm balance potentiometer is properly adjusted, there will be no rf output without an audio signal. When an audio signal is applied, the differential transistors are unbalanced and a double-sideband suppressed-carrier signal appears at the output.

product detector

The product detector circuit shown in **fig. 32** is another application of the balanced mixer. Ssb drive and bfo injection can be altered for minimum harmonic distortion; overdrive from either source results in third-harmonic distortion of the detected signal. In typical operation, the bfo voltage is 0.5 volts rms, signal voltage is 4 mV rms, third-harmonic distortion is 54 dB down and second, fourth and fifth harmonics are more than 60 dB down.

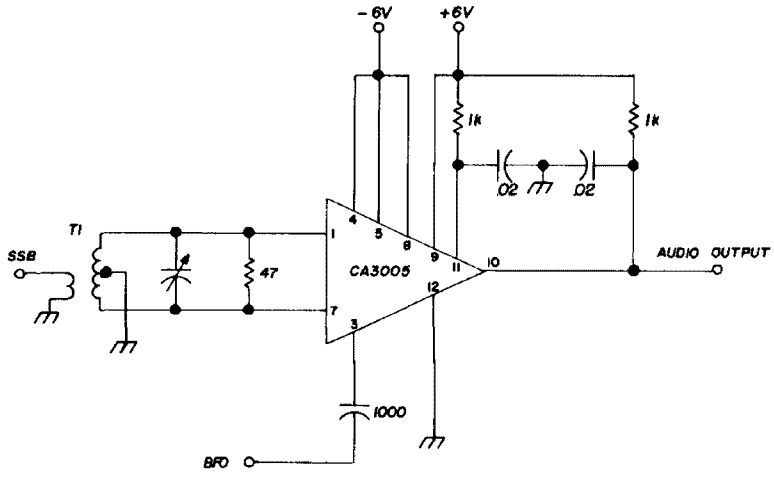


fig. 32. This product detector circuit provides excellent linearity with minimum distortion.

low-level modulator

The fine agc characteristics of devices like the HEP590 make them excellent choices for low-level a-m modulators (see **fig. 33**). The modulated rf signal can be followed by linear amplifier stages to build the signal up to several watts. The 3.5 volts bias sets operation of the modulator at its most linear point with 90% modulation along with good up-and-down modulation characteristics and very low distortion. For operation on 50 MHz, primary of transformer T1 is 6 turns number 22 AWG on a T12-2 toroid; secondary is 19 turns number 22. Primary of transformer T2 is 30 turns number 22 AWG on a T12-2 toroid core; secondary is 3 turns number 22.

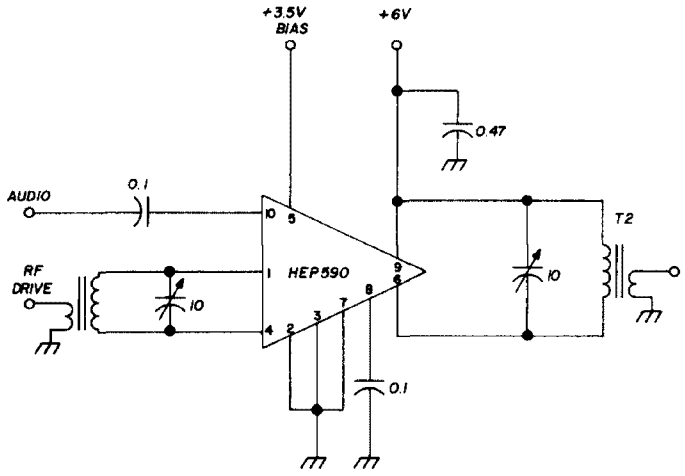


fig. 33. Low-level a-m modulator may be used in signal generators or with linear amplifier stages for communications.

oscillators

If you will remember from your basic theory, two things are necessary for oscillation: unity circuit gain and zero (or 360°) phase shift from input to output. Most linear integrated circuits (as well as many digital types) fulfill both these requirements. In fact, as you use integrated circuits in various projects, you'll find that many times you will have problems with unwanted oscillations—particularly vhf parasitics.

The oscillator circuits that are presented here are only a sample of the many circuits that can be used. However, some of them are rather unique and probably have never been used in amateur gear before so they warrant further investigation.

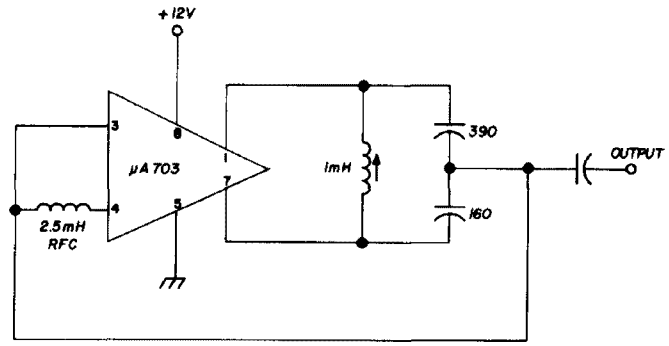


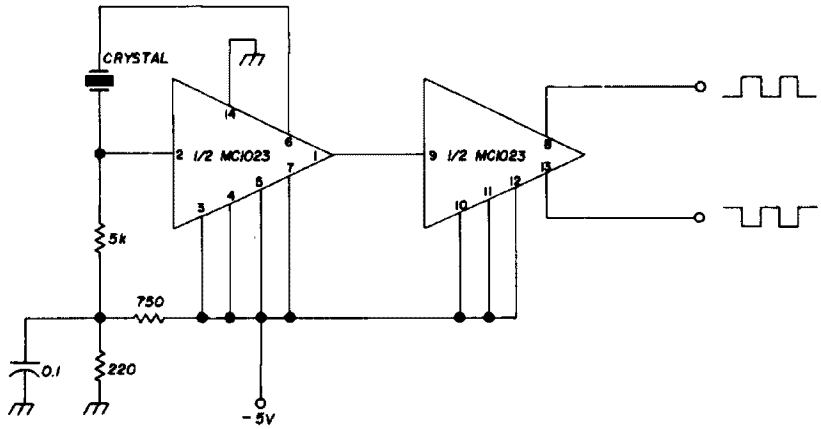
fig. 34. This 455-kHz beat-frequency oscillator may be used up to 150 MHz by changing the tuned-circuit values.

The oscillator shown in **fig. 34** is designed for use as a bfo and is very similar to the Colpitts oscillator used in vacuum-tube and transistor circuitry. Although the tuned-circuit values shown are for 455 kHz, the basic circuit may be used up to 150 MHz by simply changing the inductor and two capacitors. The rf choke between pins 3 and 4 is between the bases of the differential transistors at the input of the IC; it provides the proper bias voltage and current for symmetrical limiting.

If you look at the internal circuit of an RTL digital gate, you'll find most of the load and bias resistors for a conventional amplifier stage. If you use the correct pins, these low-cost IC's can be used for amplifiers or oscillators. The circuit shown in **fig. 35** can be used as a crystal oscillator up to about 2 MHz. Most drift in this circuit is due to crystal changes with temperature; supply voltage variations up to ±20% have a negligible effect on frequency.

Digital integrated-circuit gates that use emitter-coupled logic (ECL) provide excellent high-frequency (and vhf) response and can be used to advantage in crystal-controlled oscillator circuits. A second gate can be used as a buffer and waveshaper.

fig. 36. This crystal oscillator uses emitter-coupled logic IC's and will work up to about 20 MHz.



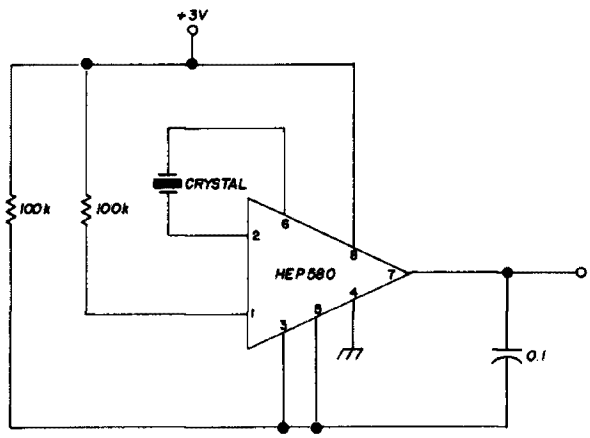


fig. 35. Simple crystal oscillator circuit is useful up to about 2 MHz.

The circuit shown in **fig. 36** is designed for crystals in the series resonant mode from 1 to 20 MHz. The differential-connected transistors in the first half of the IC operate linearly and provide loop gain while the emitter-follower outputs drive the crystal and buffer stage. In addition to acting as a buffer, the output stage works as a waveshaper; rise and fall times of the output square wave are on the order of 2 nanoseconds.

At frequencies above about 10 MHz overtone crystals are usually used. Overtone oscillator circuits are somewhat more complex than circuits designed for the fundamental because the circuit must operate at the desired harmonic. The circuit shown in **fig. 37** is designed for overtone crystals from about 50 to 150 MHz and is a modification of the circuit in **fig. 36.**

This circuit uses a tank circuit at the input to the ECL gate to insure operation at the desired overtone. At other frequencies it acts as a low impedance shunt. The variable capacitor from pin 6

to ground may be necessary to provide the proper phase relationship in the feedback loop. The second gate again serves as a buffer and waveshaper.

The oscillator circuits shown in **fig. 36** and **37** are very stable, and temperature drift characteristics are determined by the crystals that are used; voltage changes of 20% produce no frequency drift. Be particularly careful when laying these circuits out—emitter-coupled logic has extremely good vhf response so keep all the leads as short as possible and run all the unused inputs to the minus voltage supply.

The circuit in **fig. 38** uses a low-cost dc-amplifier IC—the RCA CA3000—as a crystal-controlled oscillator stage that will work up to about 2 MHz. The output waveform of this circuit has high harmonic content; to provide a sinusoidal output, put a 10K potentiometer and 0.01-μF capacitor in series with the crystal. With this "feedback adjust" control, the output waveform can be made into a pure sine wave. The frequency response of the circuit can be increased to about 10 MHz by adding a tuned circuit resonant at the output frequency; this is shown by the dotted lines.

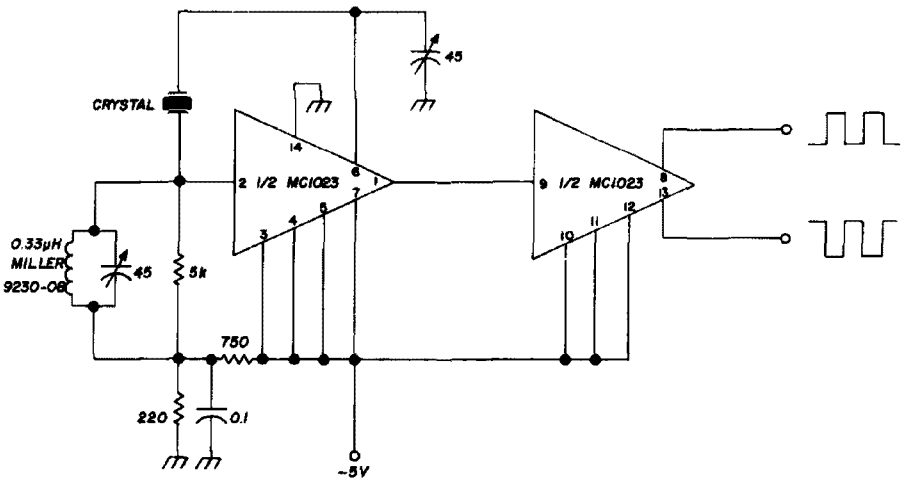


fig. 37. This crystal oscillator will work with overtone crystals up to about 150 MHz. The tank circuit is resonated to the desired output frequency.

The circuit shown in **fig. 39** uses two low-cost RTL digital integrated circuits in a crystal-controlled frequency standard with outputs at both 100 kHz and 50 kHz. The oscillator itself is the same as the one shown in **fig. 35** and uses the two gates of a HEP580. The HEP583 JK flip-flop divides the output of the crystal oscillator by two to provide 50-kHz markers. Other markers can be generated by using different divider arrangements—two flip-flops can be wired to divide-by-4 for 25 kHz markers (or another HEP583 added to the one shown in **fig. 39**). A divide-by-5 circuit for 10-kHz markers requires three JK flip-flops while a divide-by-10 circuit for 5-kHz markers requires four.

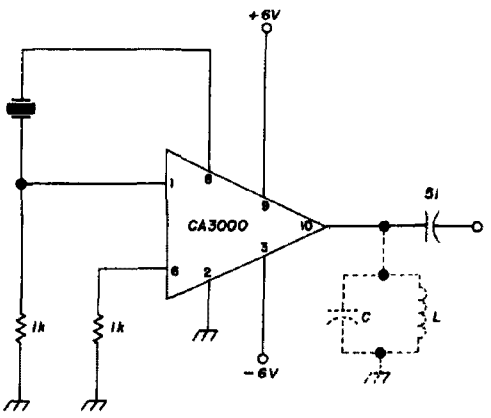


fig. 38. Simple crystal oscillator will work up to 2 MHz with no tuned circuits; tuned circuit extends range to about 10 MHz.

vfo's

The variable frequency oscillator shown in **fig. 40** may be used at any frequency between 2.5 and 13 MHz by choosing the

value of the capacitor C1. The HEP590 acts as a common-emitter, common-base amplifier when the input signal is connected to pin 1 and the output taken from pin 6; it has high gain and 180° phase shift. Therefore, to operate as an oscillator, the external circuitry must provide 180° phase shift. This can be accomplished in several ways, but perhaps the easiest way is used in **fig. 40**—the coupling transformer is inverted.

This circuit can be used anywhere in the range from about 2500 kHz to 13 MHz by simply changing the value of the tuning capacitor, C1, as shown in **fig. 40.** To decrease the range, the total capacitance may be made up with a fixed mica capacitor in parallel with a variable. To tune from 5.0 to 5.5 MHz for example, capacitor C1 would consist of a 100-pF variable, a 80-pF trimmer and a 560-pF silver mica; the trimmer is used to set the frequency range covered by the variable. The output of this oscillator has very little harmonic content; output level depends on the load that it works into. With a 12-volt power supply and a 10k load, the output is 12 volts peak-to-peak. Frequency stability is consistent with other solid-state designs.

A vfo that tunes the range from 9 to 10 MHz is shown in **fig. 41;** the tuned-circuit values can be changed to cover other frequencies in the range from 1 MHz to 150 MHz. The output level is 0.5 volts peak-to-peak, and the waveform is relatively free of harmonics. Inductors L1 and L2 are each 10 turns number 24, bifilar wound on a T44-10 toroid core; L3 is 5 turns number 24 wound over L1 and L2.

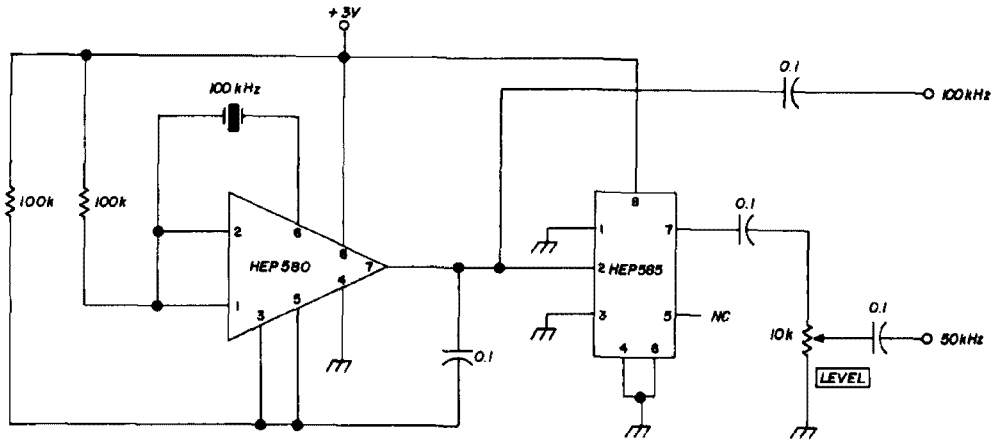
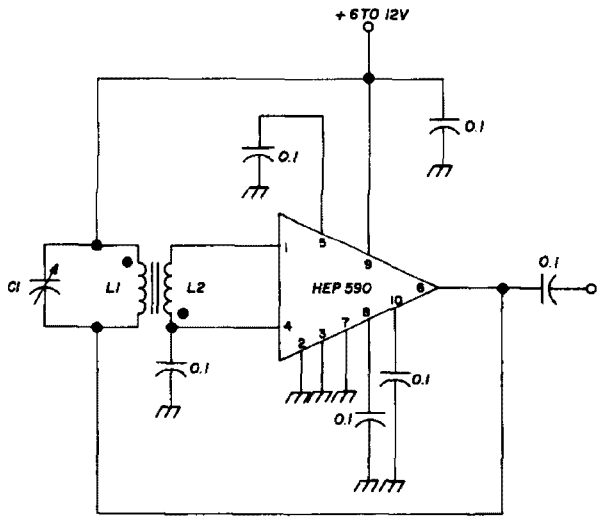


fig. 39. Frequency standard circuit provides 50-kHz markers. Output level potentiometer may not be required for some applications.



Frequency	C1 (pF)	Elmenco part. no.
2.5 - 5	780 - 2110	311
5 - 10	170 - 780	469
8 - 13	80 - 480	466

***fig.* 40. Wide range vfo. L1 is 21 turns number 36 on a T12-2 toroid core; L2 is 7 turns number 36 on the same core. Elmenco trimmer capacitors are available from Allied Radio.**

voltage-controlled oscillator

The unusual oscillator circuit shown in **fig. 42** deserves further amateur experimentation. This circuit uses two cross-coupled transistor-transistor logic (TTL or T^2L) gates to form a voltage-controlled oscillator. The oscillation frequency is determined by the cross-coupling capacitance and the supply voltage; the cross-coupling capacitance consists of the two capacitors plus wiring capacitance and internal device capacitance. With the component values shown in **fig. 42** the circuit will cover both the 40- and 80-meter amateur bands as the control voltage is varied from 4 to 6 volts. The frequency change with voltage is linear.

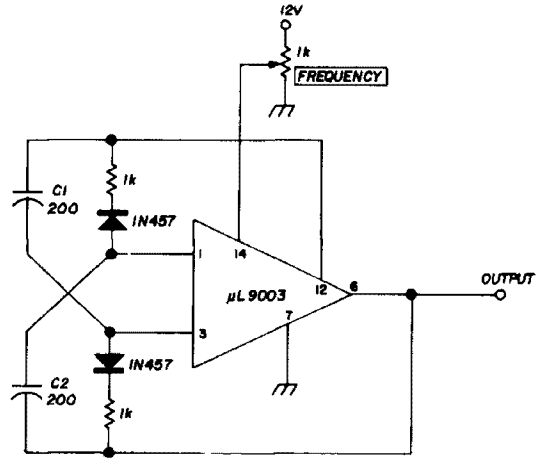


fig. 42. Voltage-controlled oscillator uses two transistor-transistor logic gates. With the components shown, the circuit covers the 80- and 40-meter bands as the voltage is varied from 4 to 6 volts.

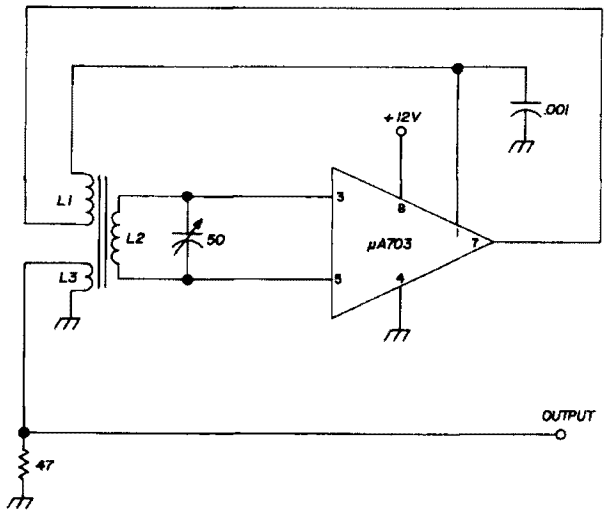


fig. 41. This 10-MHz vfo provides a clean sine wave at about 500 mV p-p.

With a stable, well filtered voltage supply this circuit can be used as a vfo with the frequency controlled by a multiturn potentiometer. The timing capacitors (C1 and C2) can be changed to cover other frequencies, although the maximum oscillation frequency is in the neighborhood of 25 to 30 MHz.

Maximum oscillation frequency is obtained with 18 to 22 pF cross-coupling capacitors although this depends upon how much stray capacitance you have in the circuit. The output waveform is a square wave but a buffer amplifier with a tuned tank circuit can be used to convert it to a sine wave.

audio oscillator

Oscillator applications of integrated circuits aren't necessarily limited to the radio frequencies. The IC Wein-bridge oscillator in **fig. 43** is designed to cover the audio frequencies although it may be used into the video frequency range. This circuit uses the μA716, an integrated circuit that provides gain options of 10, 20, 100 or 200, depending on the pin connections; for this application, a gain of ten is sufficient. The output frequency is determined by the resistors and capacitors in the positive feedback loop, and good waveform is maintained by the transistor agc circuit.

Diode D1 rectifies a portion of the output waveform and feeds it to the base of the transistor. Collector current is picked off the resistive divider, changing the voltage across diode D2. This changes the current through the diode, alters its dynamic resistance, and establishes stable operation of the IC by setting the gain at a point just sufficient to allow oscillation —this results in low distortion. The zener diode keeps the transistor from going into saturation and destroying its agc performance.

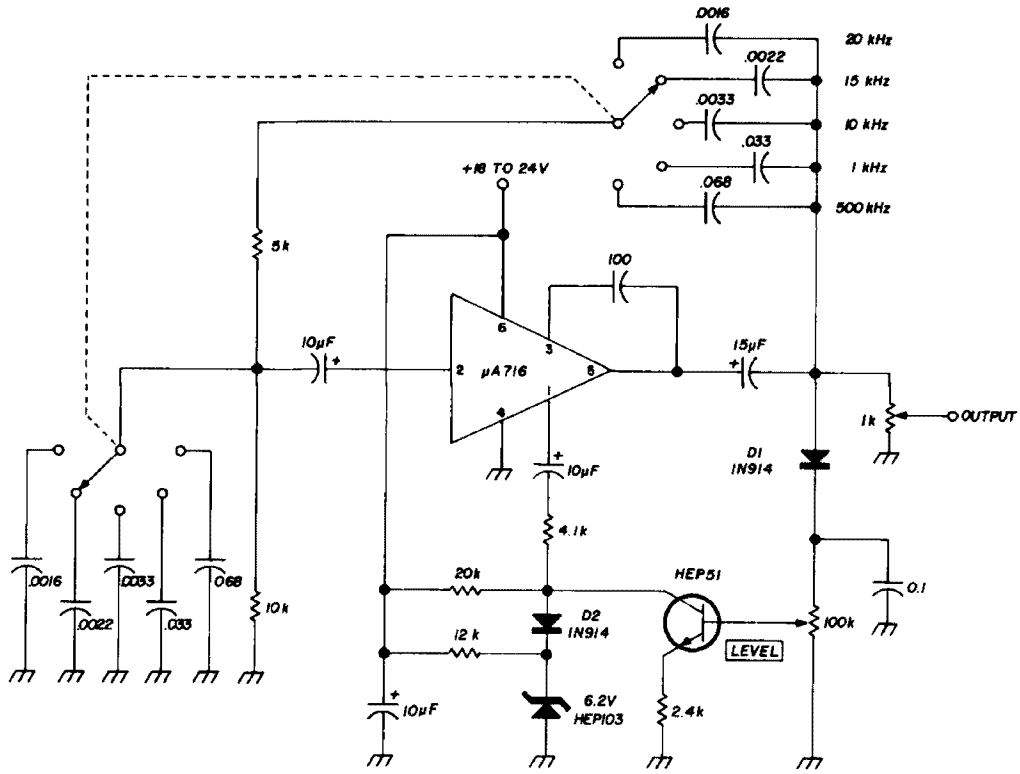


fig. 43. Audio oscillator circuit uses Wein bridge circuit to provide exceptionally pure waveform.

voltage regulators

Now that low-cost voltage regulator IC's are available, it's difficult to justify the design of regulators using discrete components. The nature of integrated-circuit construction is such that it costs very little for the manufacturer to add extra gain for improved regulation, overload and current-limiting circuits or capability for working with negative supplies. With discrete components each of these features increases cost and size.

The simple voltage regulator shown in **fig. 44** uses an integrated circuit that is designed as an audio power amplifier. However, it performs well as a voltage

Nominal Regulated Voltage	R1 ohms
10	100k
12	68k
15	51k
17	39k
20	30k
22	20k

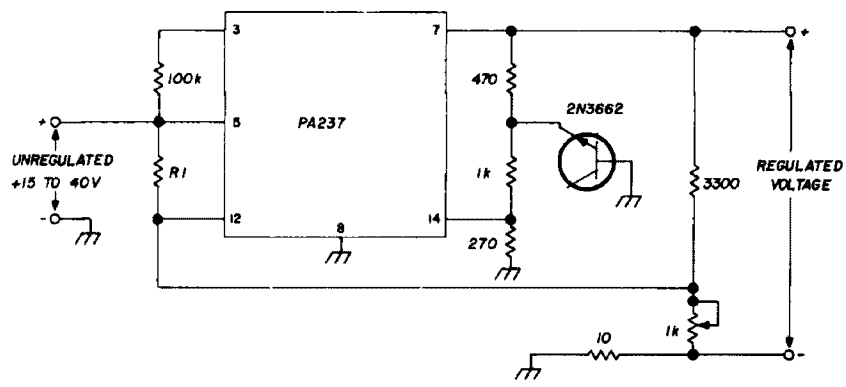


fig. 44. Simple voltage regulator using an audio power IC. Value of R1 is determined by desired output voltage as shown above. Unregulated input must be as least 3.5 volts higher than the desired output. The base-emitter junction of the 2N3662 is used as a zener diode.

regulator. Keep in mind that the PA237 is limited to 2 watts dissipation; the input current and voltage **difference** between unregulated input and regulated output must not exceed 2 watts. The unregulated input voltage can be between 15 and 40 volts; the size of R1 is determined by the desired output voltage. Short-circuit protection is provided by the 10-ohm resistor in the ground return path.

The voltage-regulator circuit shown in **fig. 45** uses an IC that was designed specifically for this purpose—National Semiconductor's LM300. The dissipation of this device is limited to about 300 mW, but an external pass transistor can be added as shown in the schematic. With the values shown here, the output can be adjusted over the range from 2 to 20 volts —depending also on the level of the unregulated input voltage. This circuit is limited to about 200 mA output with the pass transistor shown, but may be increased to 2 amperes with the circuit shown in **fig. 46.**

fig. 45. Voltage regulator provides adjustable 2 to 20 volts at 200 mA.

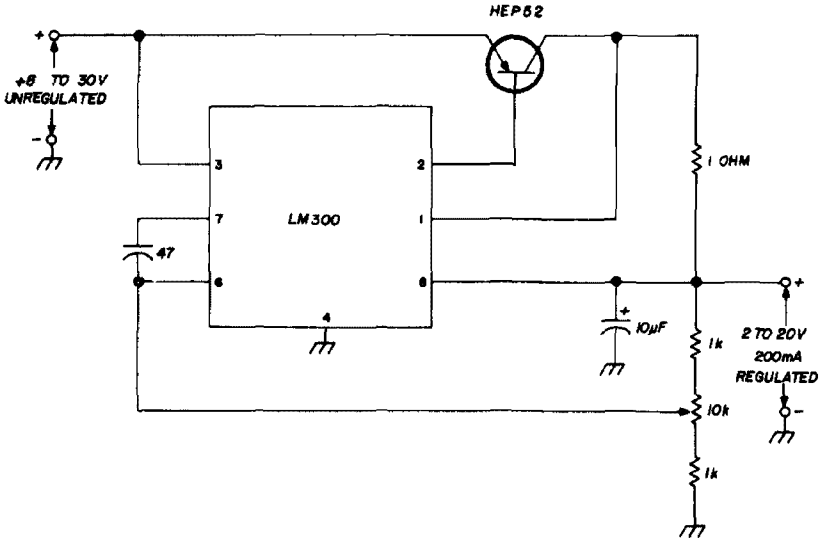
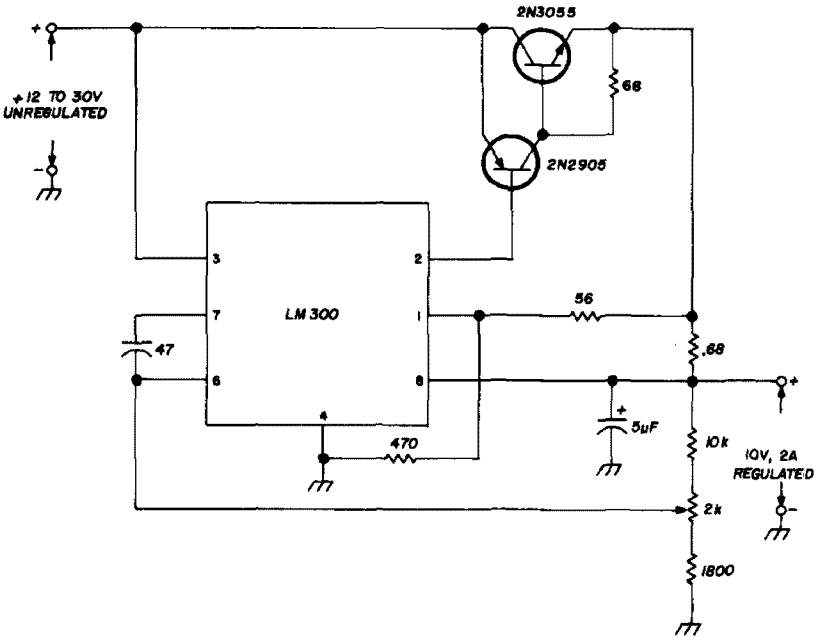


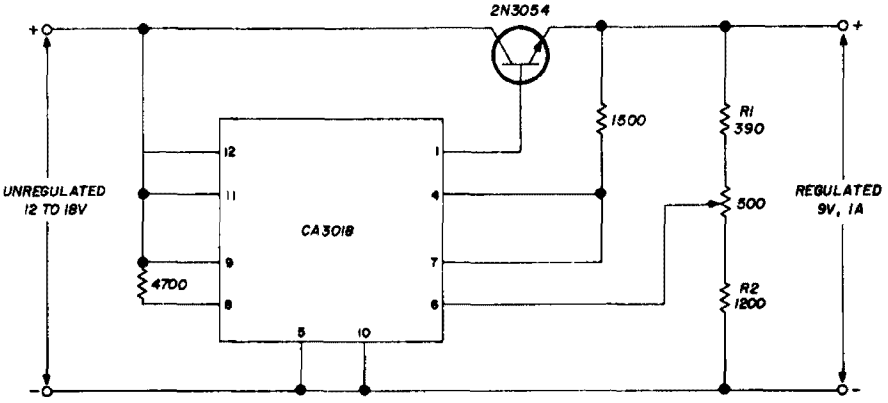
fig. 46. Current capability can be extended to 2 amperes with this circuit; nominal output voltage is 10 volts.



This circuit is designed for a nominal 10-volt output.

The voltage regulator shown in **fig. 47** will provide 9 volts output with 12 to 20 volts input. Although the CA3018 wasn't designed specifically for voltage-regulator duty, it serves admirably. The reference voltage necessary to regulator action is generated by a reverse-biased emitter-base junction. If you don't need the 1-ampere capability of this circuit, simply connect pin 1 to the output terminal, change R1 to 470 ohms and R2 to 1500 ohms—this will provide 9 volts output up to 50 mA. However, keep the input voltage low enough so device dissipation doesn't exceed 300 mW.

fig. 47. Low-cost voltage regulator uses external pass transistor to provide up to 1 ampere current capability. IC itself will handle up to 50 mA.



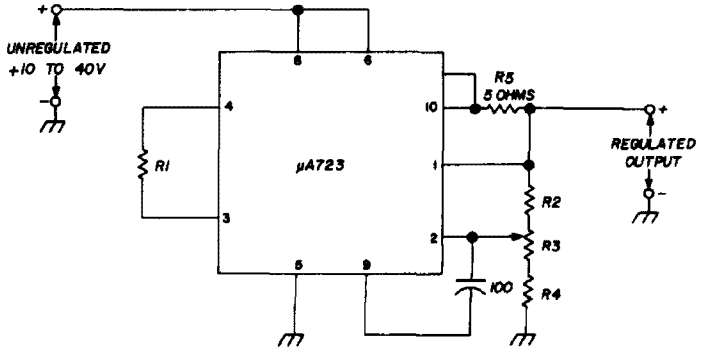
The voltage regulator circuit shown in **fig. 48** uses an integrated circuit designed for the job, the μA723. This device will handle up to 150 mA output currents by itself, or it may be used to drive higher powered pass transistors. The circuit in **fig. 48** can be tailored to your particular requirement by choosing resistors R1, R2, R3 and R4 as shown in the table. Current limiting of this regulator is determined by resistor R5, in this case in the neighborhood of 120 mA.

Typical operation of this circuit is very good. With a regulated output voltage of 15 volts, a change in input of ±3 volts results in a 1.5 mV change in output. Load changes of 50 mA result in about 4.5 mV change. The μA723 is a very versatile device and may be used to regulate negative as well as positive voltages over a very wide range. Only one application is shown in **fig. 48;** others are included on the manufacturer's data sheet.

The regulator circuit shown in **fig. 49** uses an integrated circuit that will provide up to 500 mA current into the load. Actually, the Motorola MC1460 voltage-regulator IC is available in two packages—the MC1460R in a TO-5 case that will handle 200 mA and the MC1460G in a TO-66 case that will handle up to 500 mA. The piece of silicon that constitutes the active circuit is exactly the same; package design

fig. 48. Voltage regulator provides up to 150 mA output current over a wide operating range.

Output Voltage	R1	R2	R3	R4
9	1500	750	1000	2700
12	3000	2000	1000	3000
15	3900	3300	1000	3000
28	5100	5600	1000	2000



determines dissipation. When the 500-mA version is used to drive an external pass transistor, output currents up to 10 amperes are available.

In the circuit of **fig. 49,** resistor R1 determines the output voltage level—representative values are listed in **table 3.** For output voltages other than those listed in **table 3,** the value of resistor R1 can be calculated from the following formula:

table 3. Resistors R1 and R2 (fig. 49) as a function of regulated output voltage and short-circuit current. These resistors are independent.

Nominal Regulated Voltage	R1 (ohms)	Short-circuit Current (mA)	R2 (ohms)
4.5	2k	50	13
6	4.7k	100	5.6
9	12k	150	3.9
12	18k	200	2.7
15	22k	300	1.5

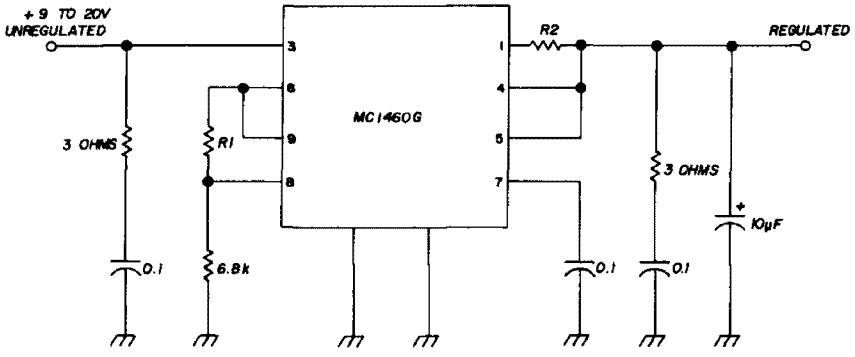


fig. 49. This excellent regulated power supply will serve most receiver needs since it will provide up to 500 mA output current. Resistor values for various output voltages are listed in table 3.

$$R1\ (\text{kilohms}) = 2(V_{out} - 3.5)$$

For an adjustable output, of course, a potentiometer could be used for R1. R2 is a current-limiting resistor; its value for various levels of short-circuit current is listed in **table 3.**

Most of the currently available voltage-regulator IC's are limited to voltages below about 40 volts, but the µA723 can be operated as a "floating" regulator at much higher voltages as shown in **fig. 50.** All that is required is that the voltage across the IC must not exceed 40 volts; in this case it is limited by the 36-volt zener diode to a safe operating point. The maximum voltage and current levels are determined by the pass transistor.

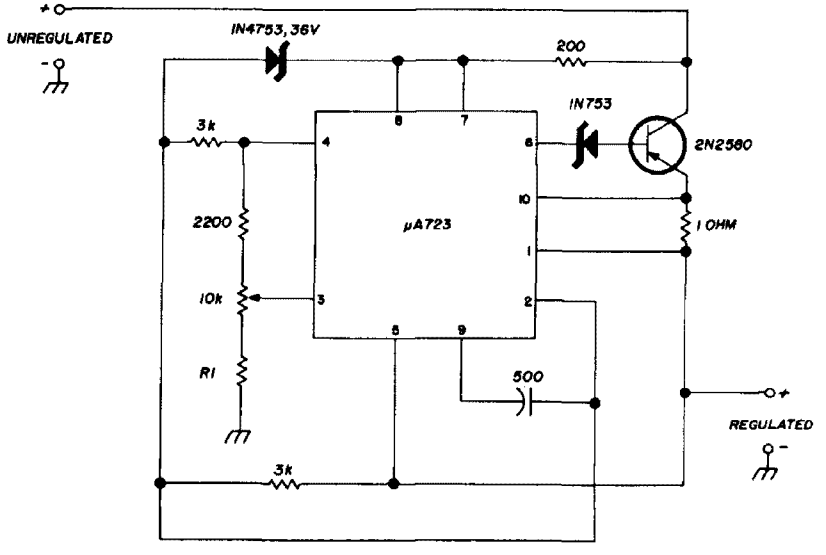


fig. 50. This circuit can be used to regulate high dc voltages. The regulated output voltage is determined by R1: R1 = 39k, E = 45 V; R1 = 68k, E = 75 V; R1 = 91k, E = 100 V; R1 = 240k, E = 250 V.

The regulation of the circuit of **fig. 50** is excellent: input voltage changes of ±20 volts result in 15 mV change in the output voltage, and 50 mA variations in load current result in 20 mV change. This device may also be used in a slightly different "floating" circuit to regulate high negative voltages; see the manufacturer's data sheet for complete details.

Many circuits—including voltage regulators—require a precise reference voltage. The circuit shown in **fig. 51** provides an adjustable reference voltage over the range of −5 to +5 volts. This circuit is particularly useful where low-level reference voltages are required. The integrated circuit isolates the zener so diode current doesn't change with the load. Since the output voltage is controlled by the potentiometer it can be set to the desired level.

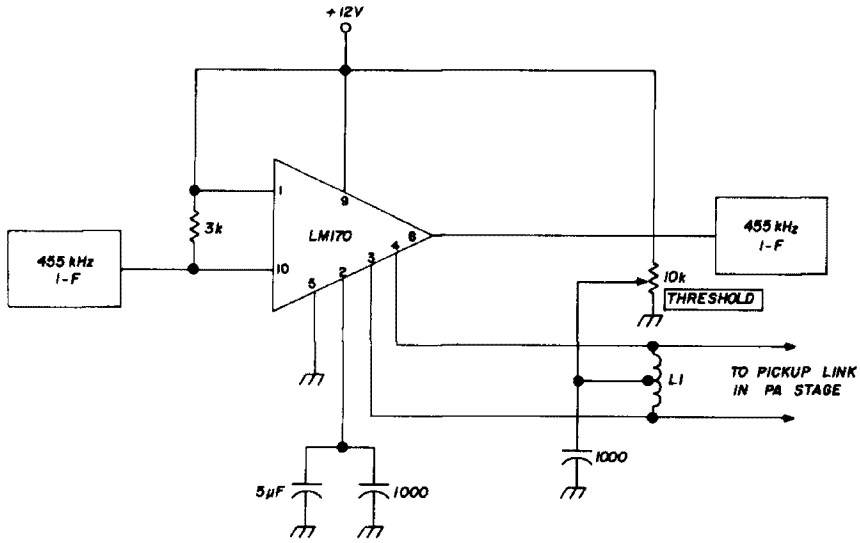


fig. 52. Alc circuit provides gain control proportional to the rf envelope of the transmitter.

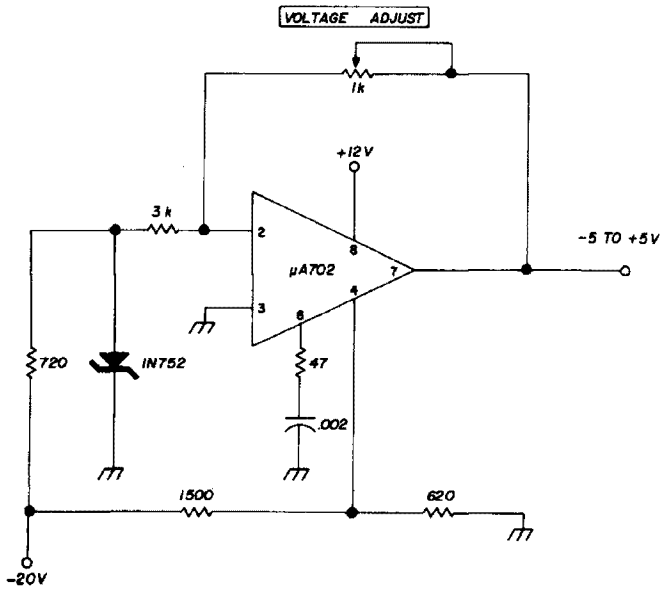


fig. 51. IC amplifier circuit isolates the zener from the load and provides an adjustable reference voltage.

other applications

In addition to the many circuits for audio, rf, i-f and voltage regulation, there are several communications-oriented applications that don't fit neatly into one category or another. One of these circuits is the alc circuit for ssb transmitters shown in **fig. 52.** This circuit compensates for variations in load impedance, tuning and supply voltage. The amplifier responds to the rf output envelope and provides control proportional to rf output. The push-pull coupling link responds to both positive and negative peaks; the threshold control can be used as a carrier level control in the absence of modulation.

The dc voltmeter shown in **fig. 53** is another handy application for integrated circuits. Most high quality commercial voltmeters use a balanced circuit but it's difficult to design a direct-coupled unit that works well. By properly selecting an

integrated circuit, you will have a well balanced circuit with high gain that does an excellent job in a voltmeter circuit. The circuit shown in **fig. 53** for example has an input resistance of 200,000 ohms per volt and is very stable. It requires no zero adjustment and has four ranges up to 1000 volts that may be read on a 1-mA meter.

To calibrate the unit, set the range switch on the 10-volt range, connect 10 volts across the input terminals and set the 1000-ohm calibration control so the meter reads full scale. The calibration control can be located out of the way since it requires very infrequent attention.

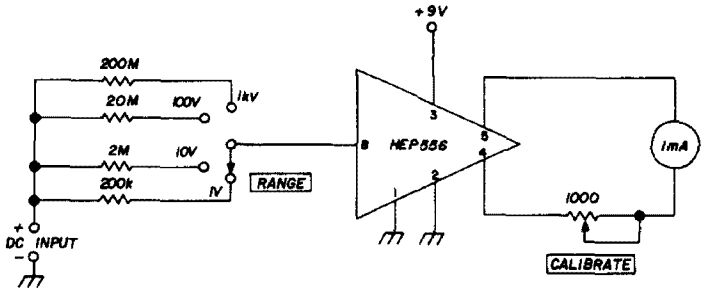


fig. 53. This dc voltmeter is very stable and linear; input resistance is 200,000 ohms per volt. The 200-megohm resistor consists of ten 20-meg resistors connected in series.

The internal circuitry of most integrated circuits is quite complex, and they will work in many more applications than those shown here. If you're interested in a particular IC, write to the manufacturer for a data sheet and a list of stocking distributors or sales representatives. Allied Radio stocks integrated circuits made by General Electric, Motorola, RCA and Sylvania. However, many major cities have distributors so it doesn't hurt to find the one nearest you. In most cases you can determine the manufacturer of a particular IC by the letter prefix on the part number: CA, RCA; LM, National Semiconductor; MC, Motorola; PA, General Electric; μA and μL, Fairchild. Here are the manufacturers' addresses:

Amelco Semiconductor, 1300 Terra Bella Avenue, Mountain View, California 94040

Amperex Electronic Corporation, Slattersville, Rhode Island 02876

Fairchild Semiconductor, 313 Fairchild Drive, Mountain View, California 94040

General Electric Company, Semiconductor Products Department, 7 Electronic Park, Syracuse, New York 13201

Motorola Semiconductor Products, Inc., 5005 East McDowell Road, Phoenix, Arizona 85008

National Semiconductor, 2975 San Ysidro Way, Santa Clara, California 95051

RCA, 415 South 5th Street, Harrison, New Jersey 07029

Signetics Corporation, 811 East Arques Avenue, Sunnyvale, California 94086

Sylvania Electric Products, Inc., Semiconductor Division, 1100 Main Steet, Buffalo, New York 14209

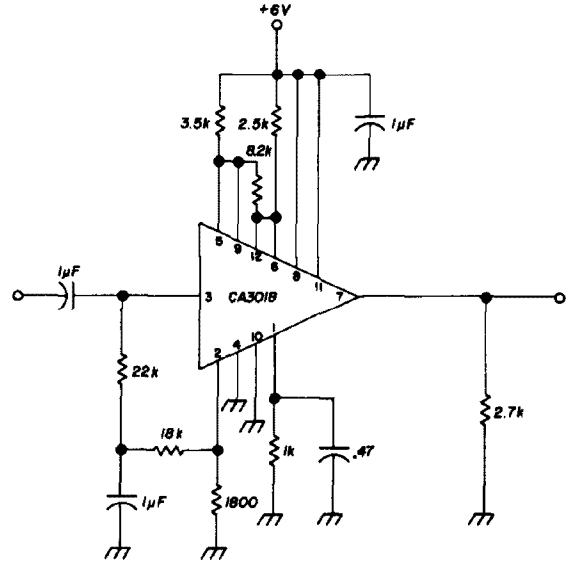


fig. 54. Broadband video amplifier covers all of the amateur bands up to 30 MHz. Mid-frequency gain is nearly 50 dB.

In addition to integrated circuits that are designed for a particular job (or set of jobs), some manufacturers are offering arrays of transistors that are on a single chip of silicon. Since the devices are physically close, their electrical characteristics are very similar; drift characteristics are nearly identical because of the high thermal conductivity of the silicon. These arrays can be used to advantage in a number of circuits where large resistors, capacitors or inductors must be used externally.

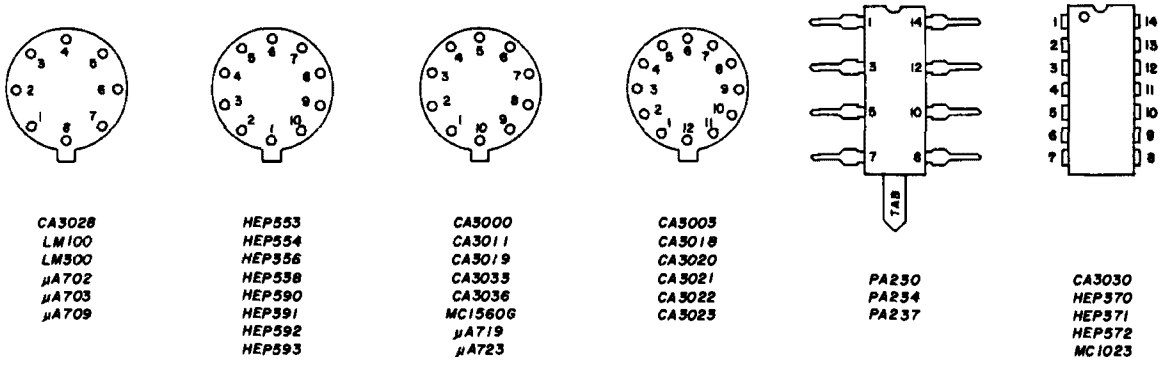


fig. 55. Basing diagrams of the integrated circuits used in this article. The circular pin layouts are bottom views; the 14-pin dual-inline packages are top views.

The RCA CA3018 transistor array contains two isolated transistors and two transistors with a common base-emitter terminal that may be used in rf amplifier service up to 100 MHz as well as video amplifiers, i-f amplifiers, class-B service or voltage regulation (see **fig. 47**). A broadband video amplifier using this device is shown in **fig. 54.**

In addition to the transistor array, RCA also offers a diode array, the CA3019, that contains one diode quad plus two isolated diodes on a single silicon chip, a dual-Darlington array, the CA3036, that contains two Darlington-connected transistor pairs and a wideband amplifier array, the CA3035, that contains three high-gain amplifier stages which may be operated individually or in cascade.

The broadband video amplifier shown in **fig. 54** is a typical application for the CA3018 transistor array. This circuit covers all of the amateur bands up to 30 MHz and provides almost 50 dB gain. With the components shown, frequency response is from 800 kHz to 32 MHz. The dynamic range is extremely good—20 μV p-p to 4 mV rms at the input.

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ham radio

receiving system degradation in fm repeaters

How to improve talk-in range using shielding and tuned-cavity filters

J. A. Murphy, K5ZBA, 1501 Maple, Apt. 18A, Lisle, Illinois 60532

The popularity of vhf fm repeaters is steadily increasing in the ranks of amateur radio despite several problems inherent in the design and operation of such stations. Perhaps the most serious and least understood is receiver blocking, or more properly, receiver desensitization.

This problem appears in two ways. First, it can severely limit the effective talk-in range of the repeater system. It can also show up as repeater chatter which is a cyclic keying of the station. First a signal breaks the receiver squelch and keys the transmitter. With the transmitter on, receiver sensitivity falls off, and the squelch closes. When the squelch closes, the transmitter shuts off, and the receiver sensitivity returns to normal; if the signal is still present, the squelch opens and the cycle repeats.

This article explores the causes and cures of receiver degradation, describes some measurement techniques, and offers some good methods to lick the problem.

noise

A transmitter will affect a nearby receiver's sensitivity in two ways. First, it can significantly reduce the receiver's front-end gain. This occurs when the transmitter carrier is present at the front end at a sufficient level to cause rectification in the amplifier or mixer input circuit. The resultant change in bias reduces the stage gain, which in turn reduces the noise input to the limiters and the limiter current. The effect of all this is ultimately to reduce the receiver's sensitivity.

The second effect is produced by the transmitter's noise spectrum. It is a sad but true fact that all transmitters produce not only a carrier and modulation sidebands, but noise sidebands as well. These sidebands may extend several hundred kilohertz on either side of the carrier. If the transmitter and receiver frequencies are only a few hundred kilohertz apart, which is usually the case in ham repeaters, the transmitter noise output that lands on the receiver frequency can be many times greater than front-end or antenna noise. This increased noise input to the limiters

produces an increase in limiter current, which in turn reduces receiver sensitivity.

While rectification decreases limiter current, transmitter noise increases it. Thus, since both effects can occur simultaneously, it's not uncommon to find them cancelling each other in regard to changing limiter current. Therefore, there may be severe receiver degradation with no apparent change in limiter current readings.

Before proceeding to a description of techniques for curing the problem, I'll develop a method for measuring the effects so corrective steps can be properly evaluated. The problem is basically how to measure receiver sensitivity, or to be more precise, the change in sensitivity. The term sensitivity refers not only to the receiver itself, but to the entire receiving system including antenna and feedline.

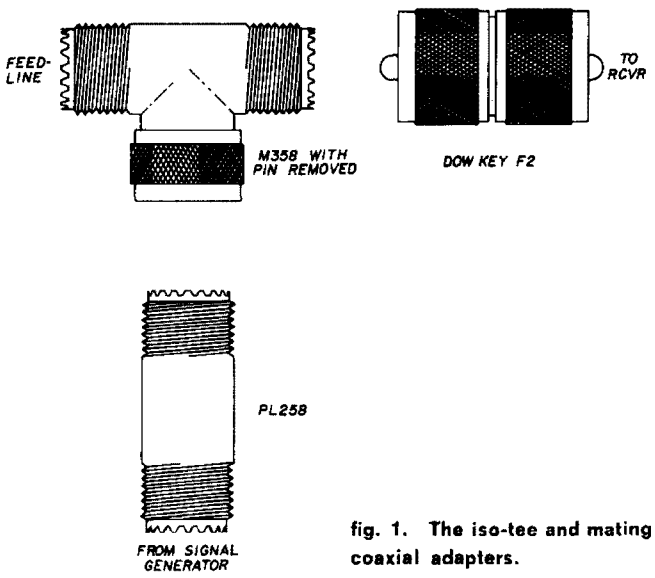


fig. 1. The iso-tee and mating coaxial adapters.

measurements

The basic piece of test gear required is a controlled signal source. A calibrated signal generator is ideal; however, most of us don't have one. Since we are primarily interested in changes of sensitity, rather than the actual measurement of it to a hundredth microvolt, simple equipment can be used with great success. You'll require a well-shielded signal source, preferably crystal controlled, and a step attenuator. Both are within the pocketbook range of most hams, supplemented by home-brewing.

Also required is an "iso-tee." This is simply a coaxial tee adapter, such as the M-358, with the center pin of the male termination removed (**fig. 1**). The female terminations are connected to the receiver and the antenna feedline. The output of the signal source is connected through the step attenuator to the male termination of the tee. The tee provides very loose coupling of the signal source to the receiving system.

receiver degradation

To measure the receiver degradation produced by the transmitter, connect an ac voltmeter across the receiver speaker terminals (see **fig. 2**). Open the receiver squelch and increase the output of the signal source until the receiver output decreases by 20 dB, just as when making a standard 20-dB quieting sensitivity check. It will probably require several hundred microvolts of rf output from the attenuator to produce this quieting, as the isolation of the tee fitting is substantial.

The next step is to key the transmitter and repeat the measurement. Ideally, it should take the same amount of signal to quiet the receiver with the transmitter on as with it off. More likely, it will require considerably more signal. If, for instance, you must inject 30 dB more signal to produce 20 dB quieting with the transmitter on, the transmitter is degrading the receiver sensitivity by 30 dB! Now the question is, what to do about it?

the cure

If the transmitter and receiver are located at the same site, the first step is to shield the two units. This includes filtering all power, control and audio leads entering or leaving the shielded enclosure. Shield kits for this purpose are available from the manufacturers, or you can brew your own. The effectiveness of the shielding can be tested by the same technique previously outlined. To test the transmitter shielding, replace the transmitter antenna with a dummy load and make the quieting measurement. To test the receiver

shielding, replace the receiving antenna with a dummy load and perform the measurements.

If the shielding is effective, no degradation will be apparent during either test. Of course, this presumes no radiation from the dummy load, so forget that light bulb nonsense. Once the units are properly shielded, reconnect the antennas and check for degradation. If it is still severe or objectionable, the next step is antenna spacing.

The transmit and receive antennas should be as far apart as possible. If they are mounted on the same tower, they should be at least 100 or more feet apart. Unfortunately, few hams can realize the benefits of 200-foot towers or separate transmitter and receiver sites, so a more practical solution is the use of tuned cavities.

tuned cavities

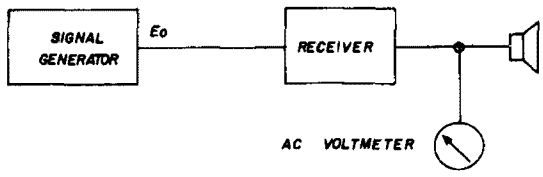
A tuned cavity is essentially a very high-Q tuned circuit. A cavity placed in the receiver feedline and tuned to the receiver frequency will pass signals on the receive frequency, while rejecting all other frequencies. A cavity placed in the transmitter feedline and tuned to the transmitter frequency will pass the transmitter signal, while rejecting noise on the receiver frequency.

The criteria for selecting a cavity, either commercial or homebrew, are physical size and power rating. As a rule of thumb, the bigger they are, the better. Many commercial units are available with a choice of coupling loops, thus providing a selection of different amounts of selectivity. There is a tradeoff involved here as the higher the selectivity, the greater the insertion loss and the lower the power rating.

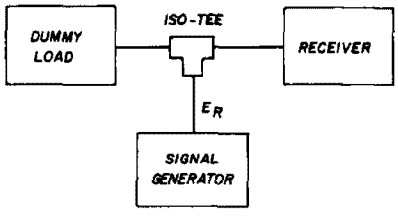
cavity location

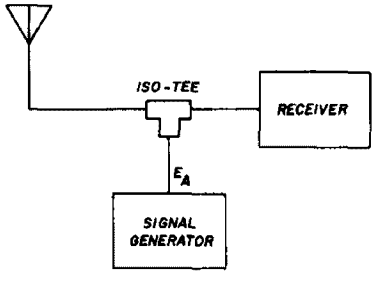
To determine the appropriate location for the cavity, watch the limiter current as the transmitter is keyed. An increase indicates degradation due to transmitter noise, and the cavity should be placed in the transmitter feedline. A decrease indicates desensitization due to rectification, and the cavity should be placed in the receiver feedline.

If the transmitter noise is the culprit and a separate exciter and final are used, try placing the cavity between the exciter and the final. Since most of the transmitter noise is generated in multiplication stages, a cavity at the output of the exciter will reduce the noise output of the amplifier. An added benefit of this location is the fact that, since you are dealing with a low-level signal, you can use a cavity with a higher selectivity (remember the trade-



BASIC SENSITIVITY = Eo = VOLTAGE FOR 20 dB QUIETING





$$\text{EFFECTIVE SENSITIVITY} = Eo\left[\frac{E_A}{E_R}\right]$$

$$\text{DEGRADATION DUE TO ANTENNA NOISE (dB)} = 20 \log \frac{E_A}{E_R}$$

$$\text{DEGRADATION DUE TO TRANSMITTER (dB)} = 20 \log \frac{E_{A\ \text{XMTR "ON"}}}{E_{A\ \text{XMTR "OFF"}}}$$

fig. 2. Test equipment setup for measuring receiving system degradation.

off) thus improving the noise reduction. Also, due to the reserve gain of the final amplifier, insertion at this point will result in less loss of radiated power than if the same filter were placed after the final.

After the first cavity is installed, a noticeable improvement should be observed. If some degradation still exists, it may be necessary to resort to additional cavities. Complete elimination of the problem may require several cavities in both the transmitter and receiver feedlines.

insertion loss

Many hams will object to the use of cavities because of their insertion loss. However, from a system viewpoint, they are the least of several ills. A repeater is useful only to stations who can both hear and be heard by it. If your repeater can be heard across the state, but it can only hear across town, then the repeater's effective range is just across town. Stations outside the talk-in range can't use the repeater and so are probably not very interested in hearing it. As far as they are concerned, it just ties up the frequency.

If you install a cavity or two on the transmitter, you may reduce your talk-out range to halfway across the state, but you may have reduced receiver desensitization to the point where you can hear halfway across the state. Thus, you've increased your effective range to halfway across the state, insertion loss and all.

Similarly, in the case of the receiver, what really counts is not just receiver sensitivity, but effective sensitivity when the transmitter is keyed. If adding cavities to the receiver feedline reduces the degradation produced by the transmitter, then you are improving performance and increasing operational range even at the expense of additional insertion losses.

receiver sensitivity

An interesting point regarding receivers arises at this point. If a receiver is capable of hearing external noise from the antenna, this noise is the limiting factor in weak signal detection. Low noise preamps and high gain antennas won't help matters, and they may produce additional problems. With an **a-m** receiver, the simplest test for sensitivity is to replace the antenna with a dummy load and see if the noise output of the speaker decreases. If it does, the receiver is hearing external noise. This test won't work with fm receivers; and if it does, you'd better start replacing tubes.

An equivalent test for fm receivers can be made by watching limiter current, but a much more accurate method is to perform the same test as used for measuring degradation due to transmitter noise. First, make the quieting check with a dummy load on the receiver. Then replace the normal antenna, and again check the quieting level. The difference in levels required to produce 20 dB of quieting is the amount of external degradation.

If the receiver sensitivity is being degraded by, say, 8 dB of antenna noise, then up to 8 dB of additional loss, either in the form of cavity insertion loss or attenuators, can be inserted in the feedline without affecting the effective sensitivity by more than 2 or 3 dB. The attenuator, by the way, will provide some improvement in desensing characteristics and considerable improvement in intermodulation protection.

If the receiver is not being degraded at all by antenna noise, then the limiting factor is the receiver front-end noise, and you may want to add a low-noise preamp or switch to a higher-gain antenna. If you decide on a preamp, be careful. Many preamps, especially those using bipolar transistors, are extremely susceptible to desensing problems. The same unit that works wonders in a mobile or base station may prove disastrous in a repeater.

in conclusion

The steps toward elimination of receiving system degradation in repeater applications are transmitter and receiver shielding, spacing between antennas and the proper use of tuned cavities. With enough cavities, it's even practical to use the same antenna simultaneously for transmitting and receiving.

ham radio



mini rtty converter

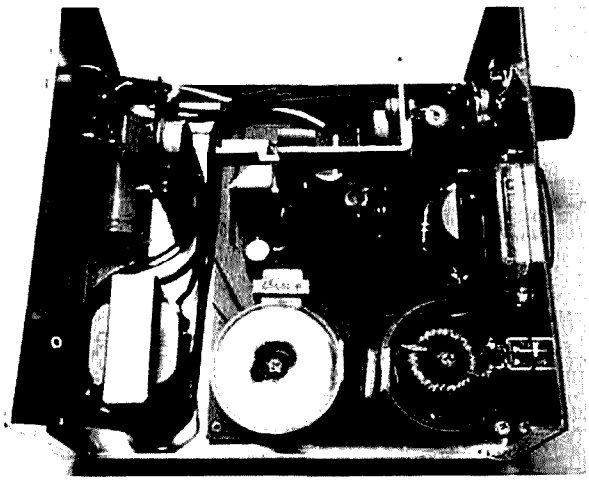
Presenting an adaptation of a standard circuit using an IC

Forrest D. Thomas, K9MRL, 3335 Central Avenue, Columbus, Indiana 47201

With the present trend toward compactness and space saving equipment, this converter should appeal to most RTTY fans.

The circuit (**fig. 1**) is not new except for the IC and has appeared in other RTTY publications. The unit is self-contained even to the power supply, and is housed in a 6x5¾x2½-inch box, which in this case was custom made.

Converter layout.



Printed circuit boards were used, and careful placement of parts is necessary to get all the components in the small package. A vertical printed circuit board was used for the IC circuit and its associated parts.

cuit. Any well-filtered, regulated circuit capable of handling the current can be used.

The integrated circuit and Q2 share a common heat sink. It is very important that heat sinks are used for long life and

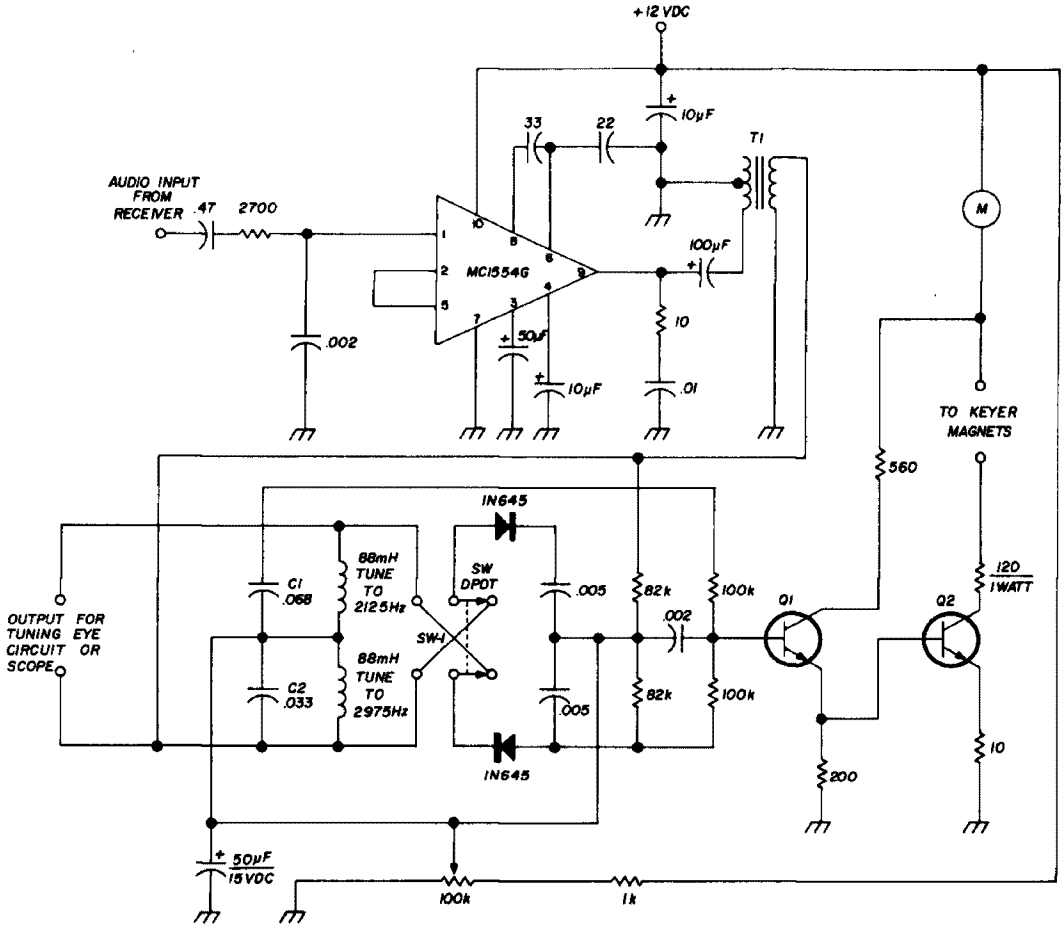


fig. 1. Schematic of the integrated-circuit RTTY converter. Transistor Q1 is a Motorola MPS6515; Q2 is a Motorola MJ3202 or RCA 40264. Transformer T1 is a 10k-to-2k interstage such as the Stancor TA35, Thordarson TR7 or Argonne AR-109. All electrolytic capacitors are rated at 15 Vdc.

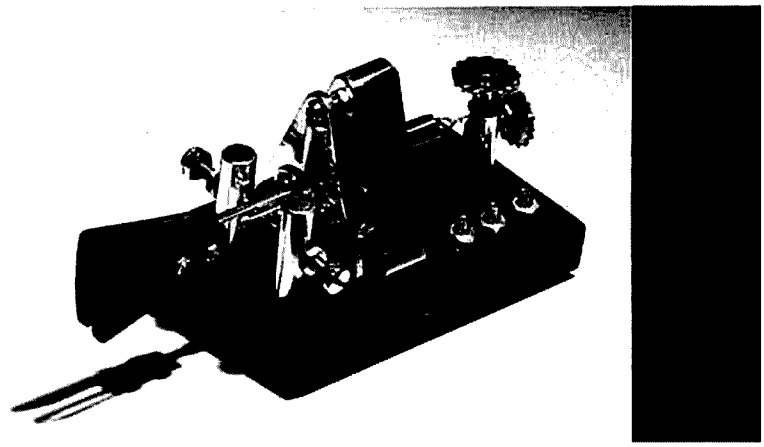
Careful tuning of the toroids is important and can be accomplished in the usual manner, using a vtvm and an audio oscillator of known accuracy. The space filter is tuned to 2975 Hz and takes approximately 0.33 µF; the mark filter is tuned to 2125 Hz and takes approximately 0.068 µF.

A regulated power supply is built into the unit, although not shown in the circuit. Any well-filtered, regulated circuit capable of handling the current can be used.

trouble-free operation. This unit has been used for some time with great success. It will give lots of trouble-free service.

The receiver input is matched to 10k as the impedance of the transformer I used was from 10k to 3.2 ohms, but in several cases the converter has been used without a transformer.

ham radio



a homebrew keyer paddle

If you love craftsmanship, here's a project to challenge your ability and patience

F. J. Chase, W3NK, 2000 Kernan Drive, Baltimore, Maryland 21207

This paddle is a sort of dare item. I was aching for a paddle to go with my newly made "TO" keyer, which I built about five years ago. So I modified my bug to a three-terminal job and began practicing on the keyer. It was rough going, because the old bug was just that—old. I acquired it second-hand from a commercial operator about 1927 or thereabouts. It wasn't new even then. I certainly can't complain about its service, but after about 40 years it was getting pretty sloppy, especially where no compensation could be made for wear and tear.

I'd visited some hams who used different types of paddles. This gave me an opportunity to try the different types. Well, you have to pay for what you get. If the price was reasonable, the thing either didn't feel right or just didn't **look** right. If it had the good qualities, then the price was out of range. So the idea of a homebrew design came into being.

the mount

The three-point rocker mount used by W8FYO seemed to be the best choice, because no sloppy bearings could develop, and any mechanical wear that did develop would

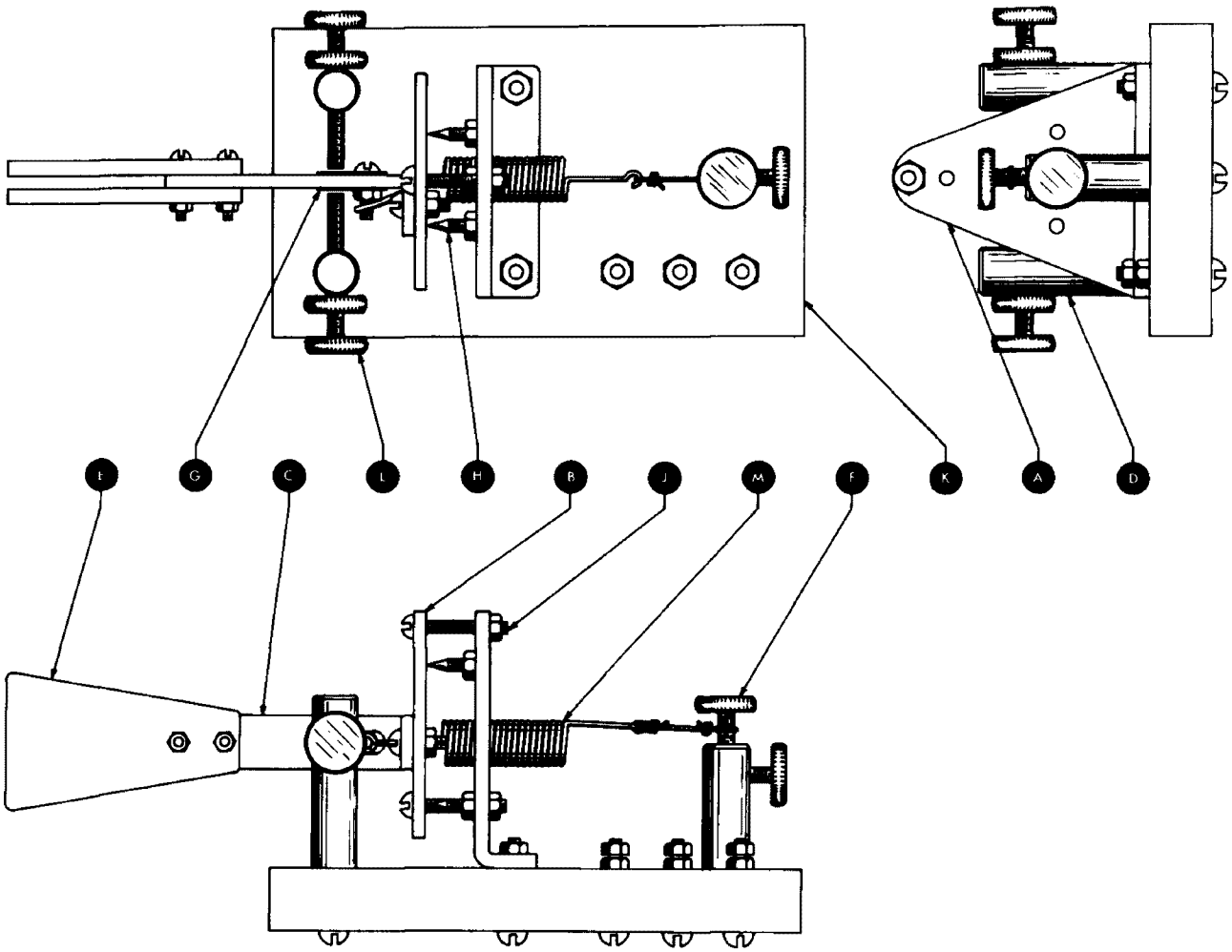


fig. 1. The assembled paddle.

be automatically taken up. It's not a copy of the W8FYO paddle, but I do wish to thank Joe for the idea of the three-point mounting.

Money aside for the moment, the main reason for the homebrew was the fun of making something with ordinary hand tools. If this design worked as anticipated, I would have a unique paddle.

performance

When completed it worked beautifully—I could do about 40 words per minute without undue strain, something that I couldn't do with the old bug for very long. But the surprise bonus was that it looked pretty good too. I took it to the office and showed it to some of my colleagues, nonham engineers. Their reaction was real joy to me. They offered lots of suggestions as to what I should do with it. The most common one was to write an article for **ham radio** magazine, so here it is. You can either make it as is, or modify it as you wish. It may even generate a bug of an idea for a design of your own.

construction

This paddle was made from scraps available in my workshop. The base is two pieces of quarter-inch Lucite cemented together with epoxy. The Lucite used to be one of those sheets that people put on their desks with notes, charts, and pictures of the family looking up from underneath. It was thrown out because it was rough and scratched so

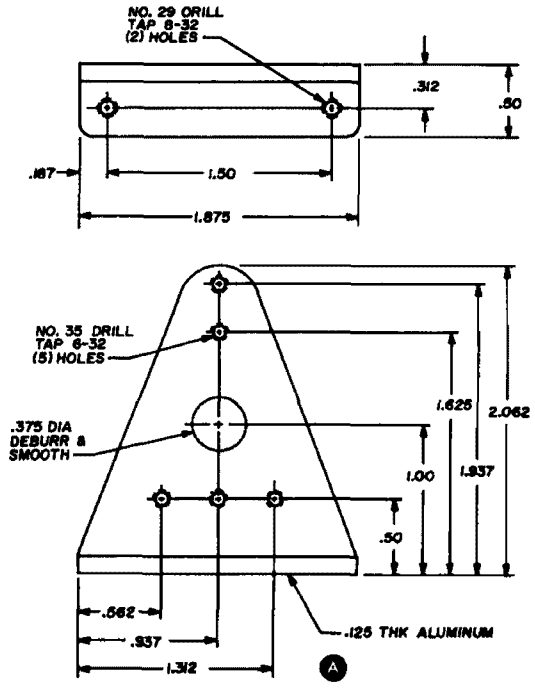


fig. 2. Main support A.

badly it lost its usefulness—except to me. However, almost any insulating material can be used such as hard rubber or bakelite.

I could have made a lead base to keep the paddle from sliding around. I decided not to, though, because heavy as my old bug was, I still had to screw it down to the desk top. So I settled for an insulating material that didn't complicate the design by having to insulate the contact posts and binding posts. One screw holds it to the desk, no matter how heavy handed you might be.

Does the paddle (**fig. 1**) look like a big undertaking? Actually it should be easier for you than for me, because I made the piece parts first, depending only on an image of what I wanted. The drawings were made after I decided to write it as a construction article. And believe me, it took much less time to build the paddle than to produce the drawings.

Except for one area, there's nothing critical despite the formidable fractional dimensions which are in inches. They're for the purists. Reasonable errors can be tolerated, because the final adjustments will compensate for any small misalignments.

the rocker panel

The critical area that requires care is marking the rocker panel to match the three points of the main support (**fig. 2**) against which the rocker panel and paddle assembly will be held by the tension spring. Precision measurements can be avoided by using the main support points as a template. But care is required. (That's the way I did it.)

After the three points have been made on the main support (A) and filed flush, carefully place the main support over the rocker panel (B) and align it so that the upper inverted-V section is matched flush. With the main support resting lightly on the rocker panel, gently tap the top screw to make a mark on the rocker panel. With the top pin

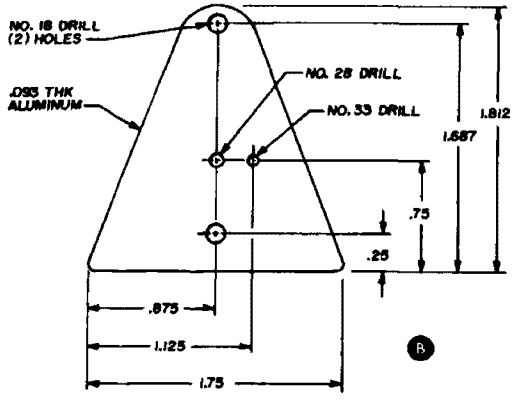


fig. 3. Rocker panel B.

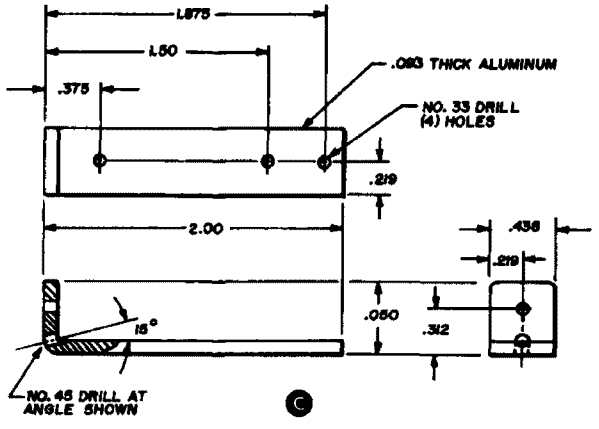


fig. 4. Paddle arm C. The wire spring feeds through the hole drilled on a 15° angle.

snugly in the mark, recheck the alignment, then tap one of the other screws. With the two screws snugly in place, tap the third screw.

the pressure points

Now that the three points have been located, they can be enlarged with a center punch. A depth of 1/32 inch is sufficient. To get the feel of how much of a wallop you'll have to apply to the center punch, try practicing on scrap aluminum.

Now that the pressure point seats have been made, a word of caution: these seats match the points on the main support. Don't change the screws on the main frame. If you have any reason to turn them after marking the rocker, chances are you won't be able to re-align the pointed screws to match the seat exactly. So be sure that the pointed screws are **set** and **tightened,** then leave them alone.

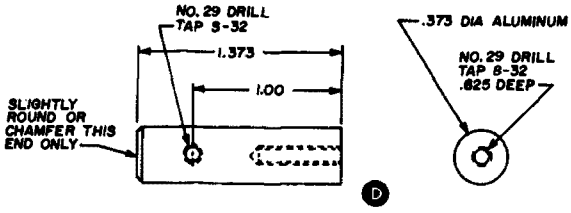


fig. 5. Terminal contact post D (two required).

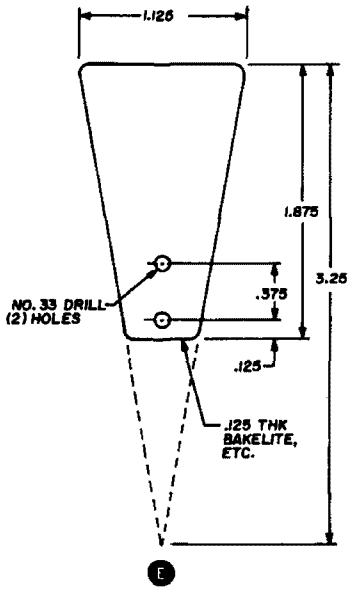


fig. 6. Paddle E (two required).

If you've been wondering why the assembly drawing looks a little different from the photograph, it's because I built two paddles. The only difference is that for the first I used a heavy chunk of aluminum for the main support, and for the second I used a piece of

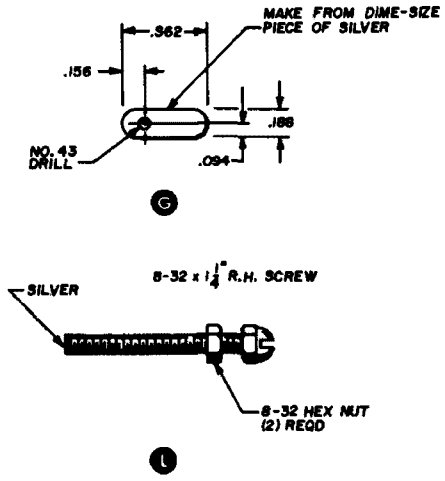


fig. 7. Silver contact G (two required) and contact screw L. Sufficient scrap should be left after making the silver contacts to make a silver contact for the contact screw. Cut a piece to size, solder to the end of the screw, and trim to size with a fine file.

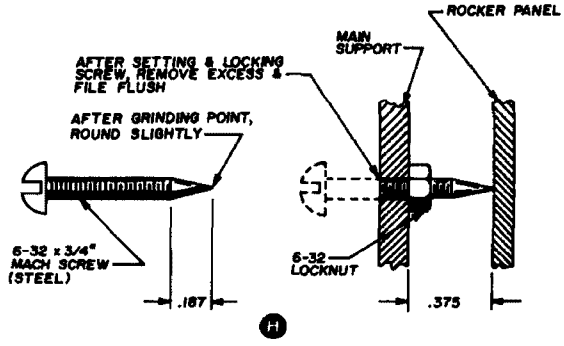


fig. 8. Support point H (three required).

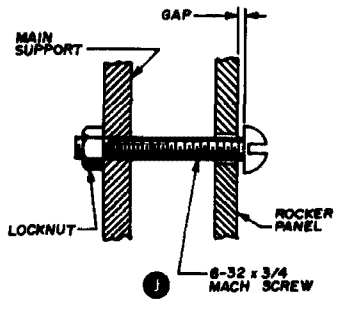


fig. 9. Retainer screw J (two required). After the rocker panel subassembly is in place, insert the retainer screws and lock with nut, permitting just enough gap to allow freedom of paddle motion.

bent, 1/8-inch stock. I made a photo of the first one and drawings of the second.

Fig. 1 shows knurled instrument-type screws and lock nuts. This was wishful thinking on my part. Actually, only ordinary round-head screws were used as can be seen in the picture. If you wish to use fancy screws, be sure that the tap matches the screw threads.

detail parts

Now to pick up the odds and ends. A 2-56 x 1/2 inch screw holds the silver contacts to the paddle arm. The screws that hold the bakelite paddle to the paddle arm are 4-40 x 1/2 inch. The terminal screws, those that hold the main support to the base, and the screws that hold the contact posts are all 8-32 x 7/8 inch. The screw that holds the paddle arm to the rocker panel is 6-32 x 3/8 inch. Before final assembly, I polished each piece part, including screws, on a buffing wheel. If you'd like to silver plate the parts, a good article on this appears in **ham radio,** December, 1968, page 62.

The second version of the paddle is presently in use at W3NK. The first version (photograph) was given to K3SFT as a memento of his visit to my station.

ham radio

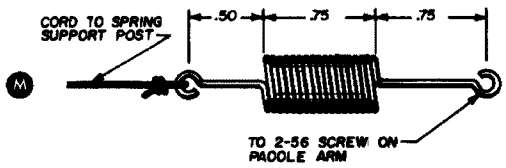


fig. 11. Tension spring M. The one I used is 3/8" diameter and about 1" long—from the junk box. Just about any small spring will work since final adjustment is made on the spring-support post (fig. 12).

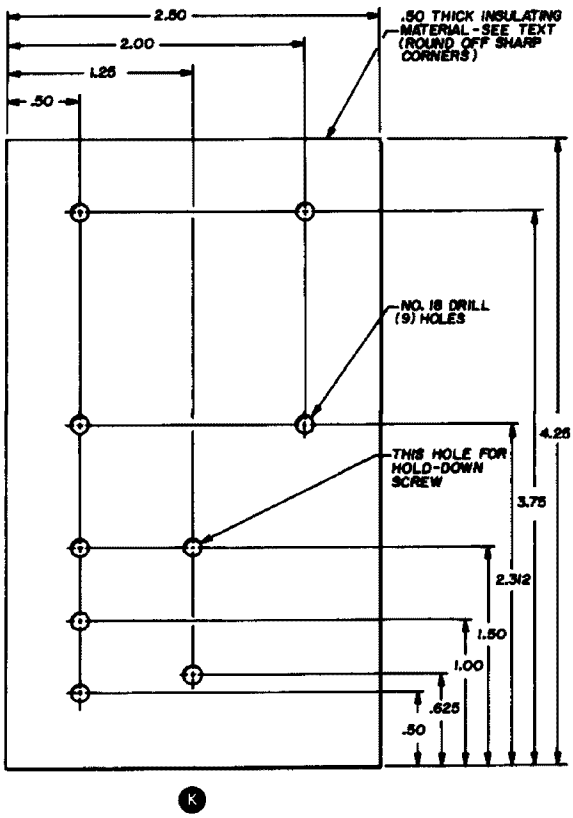


fig. 10. Keyer base K.

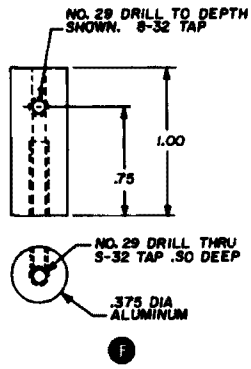


fig. 12.

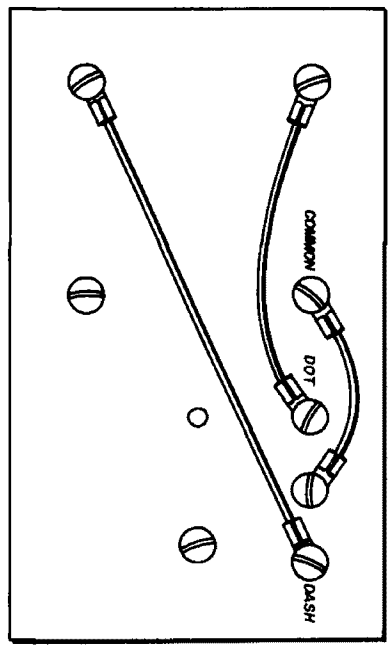


fig. 13. Base terminal wiring.

solid-state antenna switch for two meters

Adapted from a commercial design, this rf-activated switch has many features that will appeal to vhf enthusiasts

Robert M. Brown, K2ZSQ, 5611 Middaugh Avenue, Downers Grove, Illinois 60515

The subject of antenna switching without relays was discussed in fairly down-to-earth terms in a 1968 issue of **ham radio.**[1] Much is being done in this area on a more sophisticated level, however, which should be of interest to amateurs. I refer to the "mystery" band above two meters. This is known as the business radio service (BRS) band and occupies 150 to 174 MHz. Design engineers have been producing for this service some interesting circuits that are adaptable to amateur use, especially since the BRS frequencies are so close to the two-meter band.

Many manufacturers are now turning out fixed-frequency versions of their two-meter lines for use in the BRS band, which until recently has been an fm band. While single sideband has been designated for marine radio, the BRS equipment designers are just beginning to use a-m. All this has spurred new interest in equipment design, because a-m versus fm is a controversial subject in these communications circles, especially from a performance and reliability standpoint.

Let's take a look at a solid-state rf-activated antenna switch designed by Mr. K. W. Angel, applications engineer for RCA at Meadow Lands, Pennsylvania. This switch has several features that will appeal to amateurs: low insertion loss, fast response, reliability and low-cost components.

The switch exhibits about 0.2 dB loss during transmission and provides about 25 dB isolation of the receiver. As reported by Mr. Angel,[2] "Receiver sensitivity is not measurably degraded by resistive losses within the . . . diode." Furthermore, the simple device (see **fig. 4** for the circuit RCA finally adopted) has had an "excellent record of service in the field. Severe customer acceptance testing has not produced a failure."

foolproof, fail-safe

At a Canadian IEEE annual meeting on vehicular communications an RCA representative stated, "Coupled with simplicity, we have developed (switching) circuitry that is foolproof and fail-safe." It was further pointed out that no adjustments are possible that could damage the receiver, certainly a feature most two-meter men can readily appreciate.

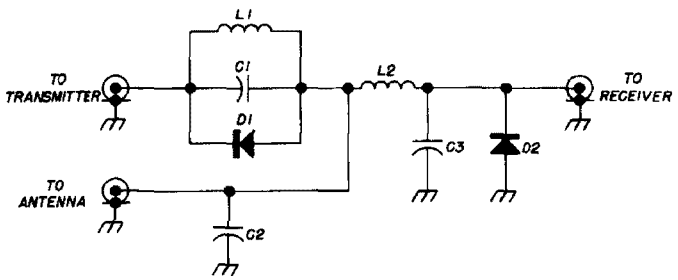


fig. 1. Essential elements of the rf-activated antenna switch. C1 and L1 are parallel resonant at the receiving frequency.

Gas discharge tubes have not been used in the past for one basic reason: poor reliability. Only with the advent of accelerated semiconductor technology has an acceptable design been possible.

RCA's design considerations included simplicity, convenience and low-power consumption—similar to conventional t-r switches. Only three external terminals would be needed: antenna, transmitter and receiver coaxial inputs (50-ohm lines). The switch could then be used with nearly all existing transceivers.

the circuit

The basic elements of the rf-activated antenna switch are shown in **fig. 1.** As the transmitter is turned off, rf signals from the antenna are applied to the receiver through the pi filter consisting of L2, C2, and C3. C1 and L1 are parallel resonant at the receiving frequency. As the transmitter is activated, the first rf-voltage cycle exceeds the diodes' breakdown voltages. This forces D1 to conduct, which shorts C1 and L1; then D2 shorts C3. What happens is shown in the equivalent circuit, **fig. 2.** Rf output is switched to the antenna, and the voltage divider, L2 and R2, isolates the receiver. C2 and L2 are selected to resonate near the transmitting frequency to avoid what the designer refers to as "reactive transmitter loading."

Damage to the receiver's first rf-amplifier transistor could occur if the circuit were left as it now stands. The reason is that the time duration, before D2 breaks down after D1 starts to conduct, could be significant. To circumvent this, further isolation of the receiver is recommended. The designer added a second filter section between D2 and the receiver (see **fig. 3**). D3 is a low-power switching diode that isolates the receiver just before D2 conducts. Interestingly, it was found that D3 turned on when it shouldn't have. The problem was resolved by placing the receiver at some multiple of one-half wavelength from the diode. In this manner, undesired signals rejected by the receiver's

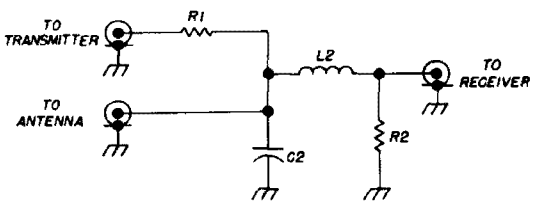


fig. 2. How the circuit appears when the transmitter has been activated. Note that C1, C3, D1, D2 and L1 have been effectively shorted out of the circuit. R1 and R2 (R2 as a voltage divider with L2) isolate the receiver from what's happening.

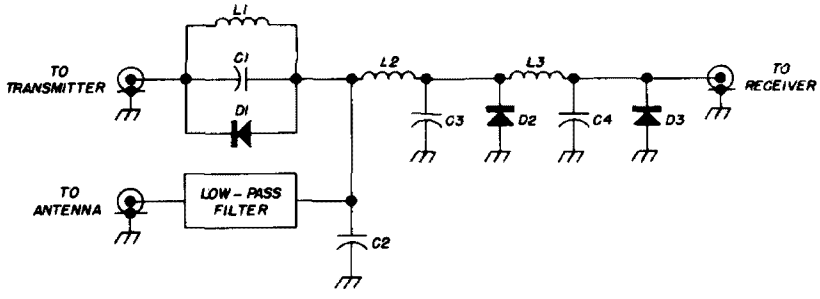


fig. 3. A second filter section has been added to prevent possible receiver damage caused by time lag between breakdown of D1 and D2. D3 doubles effectiveness of low-pass filter.

low impedance selectivity characteristics are not sufficient to cause D3 to conduct.

operation and tests

Fig. 4 represents the final circuit. Various varactors, p-n junction diodes or other switching types could be used with different forward and reverse biasing techniques. In operation, the circuit acts as a continuation of a low-pass filter inserted between receiver and antenna. On transmit, inductance L2 approaches parallel resonance with C2 and a second capacitance value on the transmit side of D1. This effectively reduces C2's value, resulting in an increase in the low-pass filter's high-end cutoff frequency. In tuneup, C2 is adjusted for minimum receiver-frequency feedthrough.

Several tests were performed that are well worth examining. To ensure that the circuit could withstand equipment faults, for example, conditions of open or short-circuited antenna were simulated. According to the designer, the most severe fault condition was run at 75° C ambient on a continuous basis for over 72 hours with no apparent change in diode performance characteristics.

From a two-meter enthusiast's viewpoint (especially mine, with my limited pocketbook), the next test was even more significant. A life test conducted at 60°C revealed that small and inexpensive glass-package diodes are fully capable of continuous operation through 174 MHz with output powers to 80 watts. At 80 watts, according to the designer, "The level of power was sufficient to darken the paint on the diode package." But no operational deterioration could be detected. Later, a larger-stud package was put into the rcuit. Power was increased from 90 through 350 watts (output). After operating for two days at the 350-watt level, no deterioration was measured.

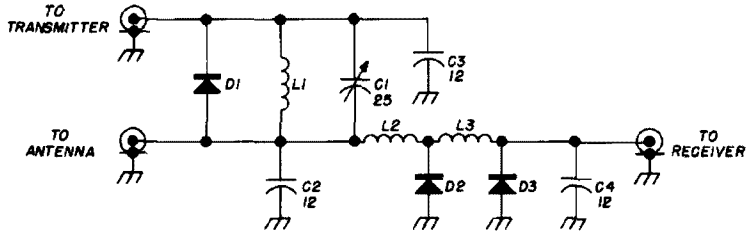


fig. 4. Final vhf antenna switch now being used extensively by communications equipment manufacturers. C1 is tuned for minimum feedthrough of the receiver frequency; it is a preset adjustment (conventional 5-25 pF trimmer) that almost never requires retuning.

Incidentally, for those of you scrambling for the soldering gun and junkbox Mr. Angel's version as adopted for use in this years line of solid-state vhf communications equipment measures a scant 1-inch high by 1-inch wide by 2¼-inches deep. And that includes the physical dimensions of the three coaxial connectors!

references

1. Robert M. Brown, K2ZSQ, "Vhf Antenna Switching without Relays," *ham radio*, September, 1968, p. 77.
2. "1966 IEEE Transactions on Vehicular Communications," *Institute of Electrical and Electronics Engineers*, April, 1966.
3. K. W. Angel, "An Rf Activated Antenna Switch," *Communications Marketing*, August, 1966.

ham radio

integrated circuit noise blanker

John J. Schultz, W2EEY, 40 Rossie Street, Mystic, Connecticut 06355

As long as we have radio receivers, noise in one form or another will probably always be with us. This unwelcome but necessary evil depends on atmospheric and ionospheric conditions, station location and the frequency covered by the receiver. In general, receiver noise is of two types: (A) short duration, high-impulse amplitude, and low repetition rate, or (B) long duration, moderate impulse amplitude, and high repetition rate. Examples of type A are ignition and click-type noise; atmospherics are typical of type B.

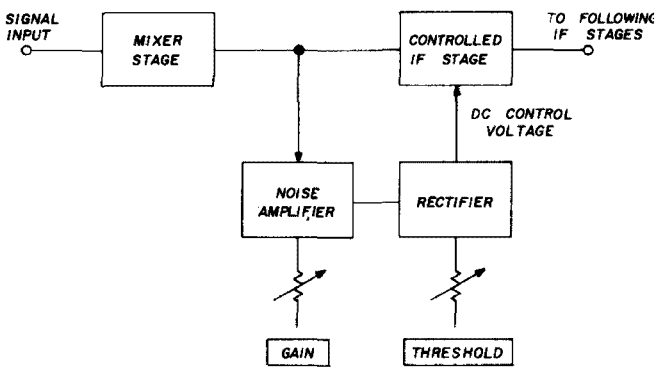
Atmospheric noise is by far the most difficult to overcome, because its characteristics are similar to those of the desired signal. Special signal processing circuits that distinguish between this type of noise and the desired signal can be designed, but they are complex and generally involve a process that compares the received signal with an "expected model" of the desired signal. Such a process is generally only possible with a digital type of transmitted signal.

There is really no need to compromise your operating by living with the racket and clatter caused by impulse noise, especially with all the technology available today using solid-state circuits that can be put together by the average amateur using low-cost devices. The following paragraphs describe a really effective noise blanker using IC's. All you need do is follow my suggestions with regard to layout and lead dress, and you can build a noise silencer that will provide many hours of operating pleasure. It could mean the difference between working that rare station and not hearing him at all.

noise suppression circuits

Many circuits have been developed to suppress the effects of high-impulse, low-repetition rate noise. Such circuits fall into

fig. 1. Basic stages of the Lamb noise blanker.



either of two categories—the first type limits the absolute value of a received signal, thus ensuring that the noise impulses don't exceed the level of the desired signal and hence aren't unduly irritating. This allows the receiving operator to perform a type of filtering wherein he distinguishes between the noise signal and the real information signal. The

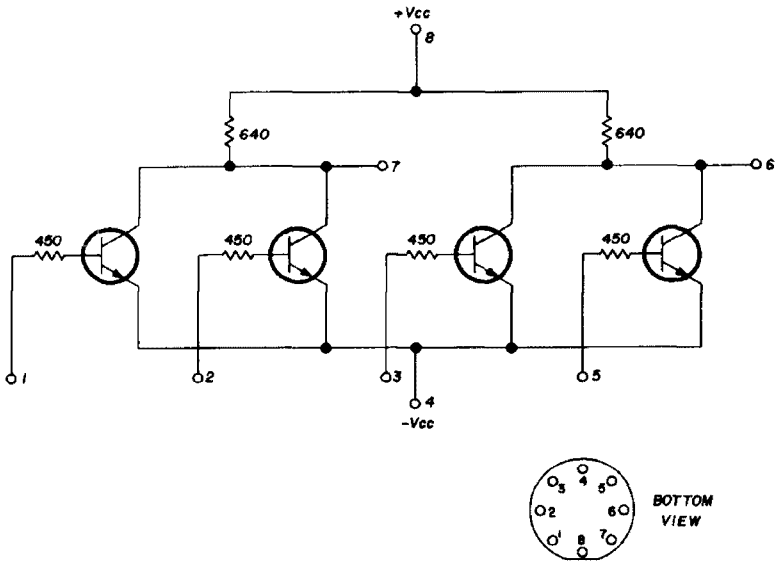
second type makes the receiver instantaneously inoperative during a noise impulse, so that the operator never hears the noise impulse and has only to integrate mentally the portions of the received signal conducted before and after the presence of the noise impulse.

a buiit-in example

This integrating process may sound rather complex, but actually it isn't. The simplest example occurs in everyday speech. Under somewhat noisy conditions, we may not distinguish every word a person says, but because we understand the language and generally the subject matter, we fill in mentally any missing portions. In fact, studies show that, for the expression of simple ideas, 30 or 40 percent of the actual expression of these ideas can be absent, and a person familiar with the language and background of the speaker will receive clear communication.

Circuits that make the receiver inoperative during a noise impulse are called noise silencers or blankers. One of the classic methods of accomplishing this action was developed by Lamb. The basic circuit (**fig. 1**) is referred to as a Lamb-type noise blanker or silencer.

fig. 2. Internal circuit of the Fairchild μL914 dual two-input gate integrated circuit used in the noise blanker.



blanker circuit operation

Fig. 1 shows that the signal at the output of the first mixer of the receiver to which the noise silencer is added is in two parts. The main portion flows, as usual, directly into the first i-f amplifier. However, a small portion of the signal flows into a so-called noise amplifier. This amplifier doesn't really distinguish between the desired signal and noise, but amplifies both noise and signal much the same as the usual i-f amplifier.

The output of the noise amplifier is rectified, and the resultant dc control voltage regulates the action of a cutoff device or circuit that makes the normal i-f amplifier inoperative. The rectifier threshold control is set so only signals of a definite amplitude from the noise amplifier will activate the cutoff circuit. Thus, the circuit can be set (by the threshold and noise amplifier gain controls) so that noise from the first mixer exceeding a given value will activate the cutoff device and control the first i-f stage output.

The basic idea of producing an instantaneous interruption in the signal output, when a noise impulse is present, can be applied to any stage in a receiver right up to the audio output stage. However, it has value only when applied as early as possible to the receiver signal-processing chain. This is because every tuned circuit a noise impulse encounters, as it passes through the receiver, tends to elongate the noise impulse due to the Q, or ringing, effect of the tuned circuits. The exact elongation a noise impulse receives depends on the receiver characteristics. However, it's not impossible for a noise impulse with a real-time duration of microseconds to be apparent as a milisecond impulse by the time it reaches the audio output stages.

Disabling the receiver for the duration of a microsecond pulse will never be noticed, but disabling it for many milliseconds may produce a noticeable decrease in received signal intelligibility. The high Q of the i-f circuits also accounts for the fact that many conventional noise limiters produce no real benefit for ssb or cw reception.

By the time a sharp noise impulse reaches the detector, so much circuit ringing and

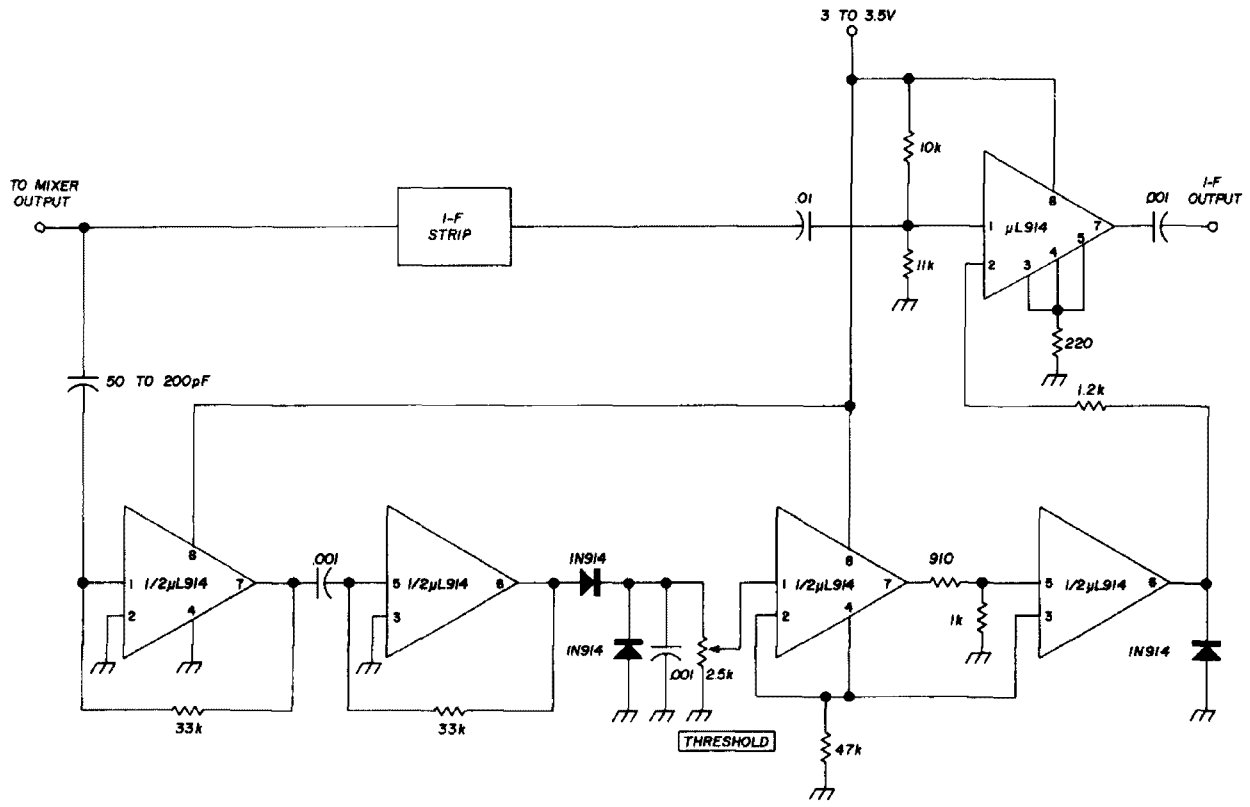


fig. 3. Schematic of the integrated-circuit noise blanker. All IC's are Fairchild μL914's or equivalent.

impulse elongation have occurred that severe signal distortion is present, which no limiter can change. So, with modern receivers, some variation of the Lamb circuit offers the only practical solution to eliminate the effects of impulse noise.

the ic lamb unit

The integrated-circuit version of the Lamb silencer uses three inexpensive Fairchild μL914 units. **Fig. 2** shows the circuit of each μL914 unit, which contains four transistor functions.

The tube-type Lamb unit used a loosely coupled i-f transformer at the input to the noise amplifier to gain some degree of signal/noise discrimination **and** a diode switch or double-grid tube such as a 6BE6 either to act upon or replace the first i-f stage to achieve cutoff. The IC uses a slightly different approach, although the over-all operation is the same as the tube unit.

Fig. 3 shows the wiring of the IC. The first μL914 on the left is a common emitter amplifier. It provides a degree of noise discrimination due to the time constants in the coupling circuits. The second IC is a Schmidt trigger. This is the main area in which the IC circuit differs from the tube circuit. The output of the trigger goes positive when the input voltage exceeds a certain value. Then the trigger switches to zero output when its input voltage falls below another set value.

The input voltage required to make the trigger switch is adjusted by a triggering level control, which is the equivalent of the threshold control in the usual Lamb circuit. The level control is also used to control simultaneously the gain of the noise amplifier.

The last μL914 unit is a simple diode switch. Positive pulses from the Schmidt trigger cause the switch to disable the i-f signal path. The fact that no rectification takes place, as in the usual Lamb circuit, means that switching is extremely fast, and no compromise is necessary between rectifier output filtering and a short time constant.

construction and adjustment

It's almost impossible to describe a unit that will universally fit any receiver. Therefore, the IC unit should be breadboarded before installation. The unit shown will work nicely with many receivers, but others may require that the signal from the detector stage be further amplified. This can be done by connecting another IC amplifier circuit similar to the first μL914 shown in **fig. 3** in series with the first amplifier.

Also, an i-f transformer can be used between the detector stage and noise amplifier to provide increased voltage gain. The circuit should be usable at i-f's to about 2 MHz. At higher frequencies some signal leakage through the last μL914 switch will be noticeable. Other dual-input gate IC's may be used in place of the μL914's if information on their biasing is available.

Construction can be accomplished by grouping the circuit components on a piece of Vector board, which is then mounted by the detector/first-i-f stage. I recommend that the unit not be used externally, because the effect of long signal cables can easily degrade performance. Care should be taken that the leads to terminals 1 and 7 of the last μL914 switch are well separated to prevent signal leakage. Also note that the μL914 switch breaks the dc line between the first i-f transformer and the base of the first i-f amplifier.

If bias for the stage is fed through the secondary of the i-f transformer, it must be rerouted to go after the coupling capacitor from terminal 7 of the last μL914. Connection of the unit will also generally require simple repeaking of the first i-f transformer primary and secondary.

The triggering level control carries only dc and may be located wherever desired; it can be an external panel control, or it can be set and left alone if continuous noise silencer action is desired.

For the reasons previously stated, the noise silencer should be placed well ahead of the receiver selective circuits. For receivers having broad selectivity, typical of some vhf types, the silencer may still prove reasonably effective if placed later in the i-f strip, or possibly even in the audio circuit.

ham radio

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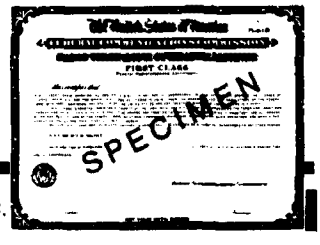
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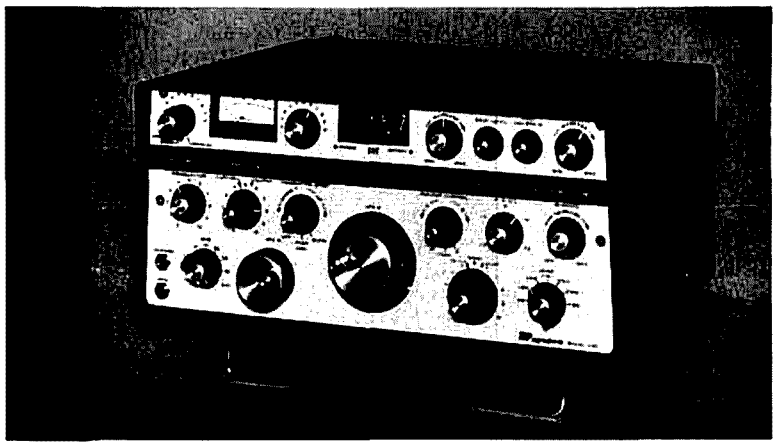
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the integrated station

A quick look at some of the design highlights of the new Signal/One amateur transceiver

Skip Tenney, W1NLB, Publisher, ham radio magazine

Nearly all of us have had our curiosity aroused by a recent series of advertisements by a new name in the amateur field, Signal/One in St. Petersburg, Florida. While I was in the Sunshine State to attend the Tropical Hamboree in Miami I decided to take an extra day and see just what was going on on the other side of the state.

I found Signal/One in a small modern plant in the shadow of their parent company, the ECI Division of NCR. A quick tour of ECI disclosed approximately 2000 people producing all types of military equipment from sophisticated satellite terminals to small man-pack sets for use in underdeveloped parts of the world.

But the visit became really interesting when Don Fowler, W4YET, project engineer for the CX7 of Signal/One and Dick Ehrhorn, W4ETO, the general manager, introduced me to their new integrated station.

It is a transceiver, whether or not their ads are willing to accept the fact, but the title "integrated station" is also a very appropriate description. Not only do they offer an excellent transceiver, they have included in the same box a 115/230-volt, 60-Hz power supply, an extra vfo, an IC keyer, an rf clipper plus a noise blanker. The complete transceiver is solid state with the exception of the final power-amplifier stage.

The standard unit covers all amateur bands between 1.8 and 29.7 MHz. A $5 crystal permits operation in one 1-MHz segment in any one of the bands from 2 to 3, 4 to 7, or 8 to 14 MHz; three spare bandswitch positions are included for this purpose. As an extra feature the CX7 offers a four-digit frequency counter with Nixie® tube readout that gives you tuning accuracy down to 100 Hz.

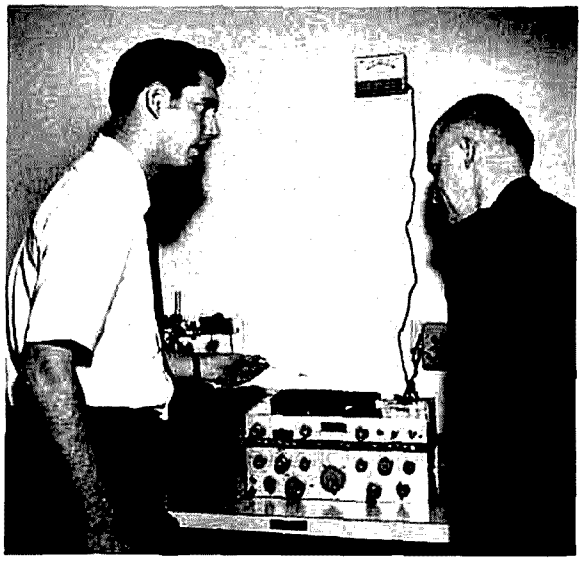
You have your choice of broadband or tunable transmitter output at the change of a switch; if your antenna has an swr below 1.5:1 it is absolutely unnecessary to tune the transmitter in any way—the receiver preselector needs peaking, of course—just dial the frequency at which you wish to operate, and start talking. The transmitter features an RCA 8072 operating at a conservative 300 watts PEP under steady-state conditions. It is conduction cooled and has more than ample heat dissipation. The alc circuitry offers both grid and screen protection assuring the longest possible tube life.

Probably the most interesting part of the CX7 is the receiver. The designers have incorporated two extremely steep-sloped filters, with a combined bandpass of 2.0 kHz at 6 dB down and 3.0 kHz at 60 dB down, together with the niftiest new passband tuning arrangement (Electronic i-f shift*) that anyone has yet to come up with. Signal/One is also promising an outstanding CW filter, but it was not in the unit which I used.

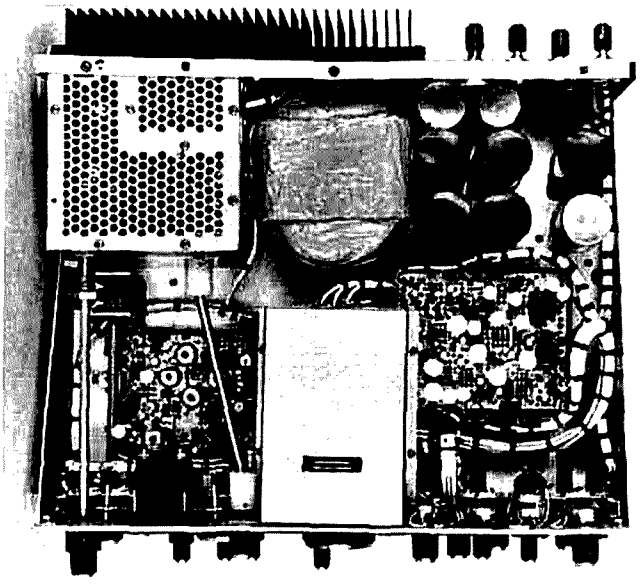
By tuning the bfo crystal oscillator (**fig. 1**) both second mixer injection and bfo injection into the product detector are varied by an equal amount. Thus, as you vary this tuning control, the signal you are listening to remains unchanged but its relationship to the i-f passband can be completely altered. With the very sharp filter skirts you can listen to a CW signal and drop an interfering signal in and out almost as though it was being switched on and off. Even a 1000 microvolt signal from a signal generator was easily eliminated —whether just above or just below the desired signal. When listening to the 40-meter phone band I found no signal that could not be comfortably copied through the European broadcast QRM that is so prevalent on the East Coast in the late afternoon.

* *patent applied for*

Don Fowler, W4YET, project engineer at Signal One, left, and Skip Tenney, W1NLB, right, with the engineering prototype of the Signal One integrated station.



Compact but uncrowded layout of the CX7.



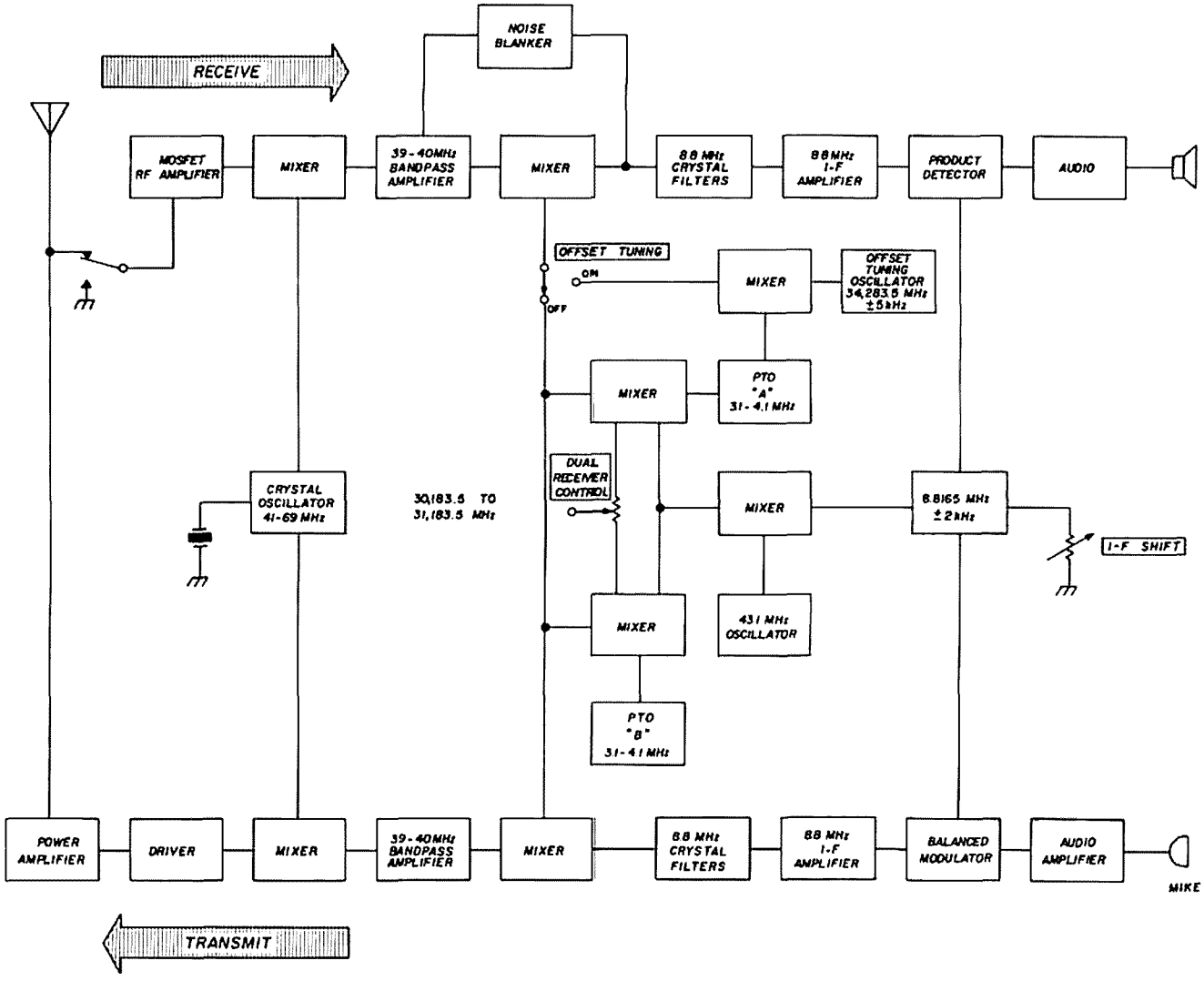
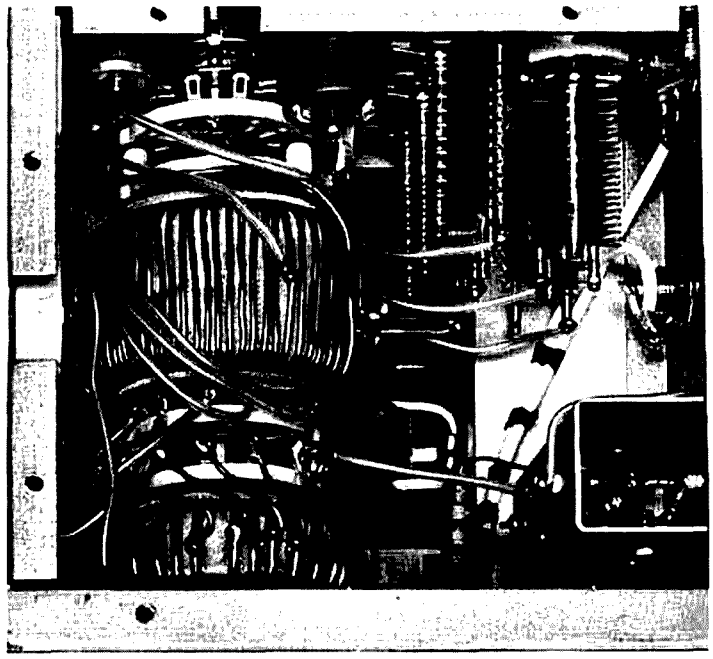


fig. 1. Block diagram of the Signal/One CX7.

Not only does this rig offer outstanding selectivity, it scores very high in the versatility department too. It can be operated as a straight transceiver offering several kHz of offset tuning, or you may listen to two channels simultaneously and transmit on either channel at the push of a button. Receiver preselector tuning is independent of the transmitter so split-frequency operation can literally be from one end of the band to the other; the broadband transmitter circuits completely eliminate any form of transmitter tuneup.

In the block diagram you can see many of the interesting features. The diagram has been simplified to eliminate much of the special switching, but all of the basic functions are shown. Special functions such as vox have been left out for clarity al-



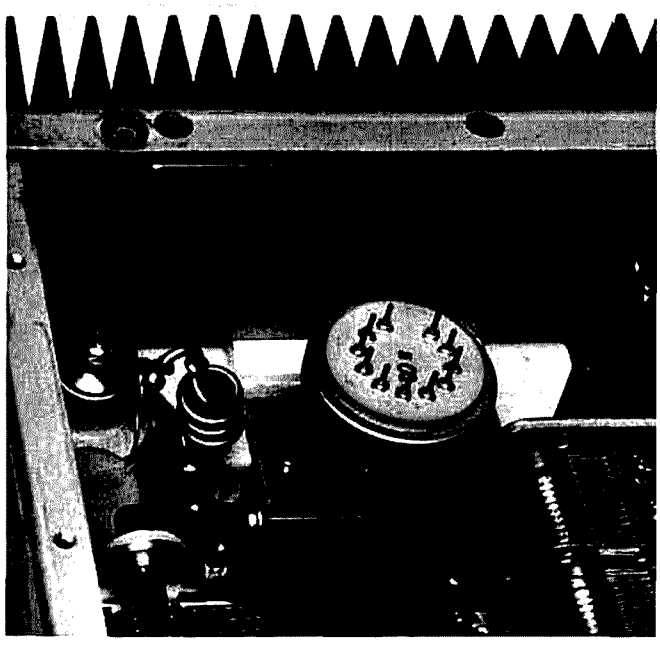
though they add much to the usefulness of the entire unit.

This design has made extensive use of the latest devices. For instance, dual-gate mosfets are used in the front end, in most mixers and in the high-frequency i-f stages. IC's are incorporated into the audio output, the electronic keyer, the frequency counter, the low-frequency i-f strip and the final broadband mixer in the transmitter.

Another interesting design innovation is the use of ground-plane circuit boards. This refers to the use of printed-circuit boards with copper foil on both sides. On the side on which the components are mounted the foil is grounded and acts in much the same manner as a chassis in conventional construction. The foil on the other side of the board is used for circuit wiring in the normal manner. Signal/One claims this offers much better stage isolation and stability.

Although there is much more that is new in this unit and well worthy of comment, I have only tried to hit the high spots of this new design. There should be a number of complete technical reviews published as the CX-7 becomes more available.

The final-amplifier tube is clamped to a block of Beryllium oxide that provides a thermal path to the heat sink as well as high-voltage insulation. The final tank circuit (left) uses a low-Q toroid design.



specifications

general

frequency coverage:	1.8 to 29.7 MHz transceive,
vfo's:	dual permeability-tuned oscillators, resetability to 100 Hz
frequency readout:	built-in frequency counter with digital readout to 100 Hz
stability:	less than 100 Hz in first half hour; less than 50 Hz in any hour thereafter at fixed ambient
CW keyer:	built-in, 5 to 50 wpm
power supply:	built-in, 115/230 volts, 60 Hz
sensitivity:	better than 10 dB signal-plus-noise-to-noise ratio with .33 μV at 10 meters (2-kHz bandwidth)
selectivity:	2 kHz at —6 dB, 3 kHz at —60 dB provided by two cascaded crystal filters; optional CW filter available
image rejection:	80 dB
i-f rejection:	60 dB
agc:	less than 6 dB audio output change for signal level from 1 μV to 100 mV; selectable hang time
i-f shift:	variable up to 2 kHz above and below normal
noise blanker:	active blanker with adjustable threshold
power:	300 watts PEP input, 150 watts minimum PEP output on all modes and bands
carrier suppression:	60 dB
unwanted sideband:	60 dB down
distortion:	third-order intermodulation 30 dB below each of two equal tonos at full-rated output
tuning controls:	none for amateur bands when load swr does not exceed 1.5:1
power amplifier:	conduction-cooled 8072 ceramic-metal tetrode
duty cycle:	continuous at full-rated input, all modes
speech processing:	rf envelope clipper built-in

propagation predictions for may

A discussion of the ionospheric E-layer, sporadic E and vhf and h-f propagation for the month

Vivtor R. Frank, WB6KAP, 12450 Skyline Boulevard, Woodside, California 94062

The ionosphere is divided into three main regions: the D region that extends from approximately 30 to 55 miles above the earth's surface, the E region from 55 to 80 miles and the F region from 80 miles to 250 miles or more above the earth's surface. This month's column will primarily be concerned with processes that occur in the E region.

The earth's atmosphere is well mixed and chemically homogeneous below the E region. At greater heights the atmospheric constituents are separated according to their atomic weight. The most important change in the neutral atmosphere in the ionosphere is the dissociation or breaking apart of diatomic oxygen (O_2) to monatomic oxygen. From heights greater than 80 miles to the peak of the ionosphere, monatomic oxygen makes up the largest part of the atmosphere; at greater heights lighter gases dominate.

At a height of about 67 miles, diatomic nitrogen (N_2) comprises 69 percent of the atmosphere (compared with 78 percent at the earth's surface). Monatomic oxygen comprises 25 percent and diatomic oxygen 6 percent of the atmosphere (compared to 21 percent at the surface of the earth).[1] In addition there are traces of argon, monatomic nitrogen, monatomic hydrogen, neon, helium, krypton, xenon, ozone and water in descending order of occurrence. There are about 2.4×10^{12} molecules per cubic centi-

meter at these heights, but less than one in ten million is ionized at any one time. But it is the coherent minute motions of trillions of free electrons under the influence of radio-frequency electric fields that makes ionospheric radio propagation possible.

Solar flux of ultraviolet and x-rays is primarily responsible for photo-ionization in the ionosphere. Photo-ionization is the dissociation of a molecule into a positive ion and a free electron by the absorption of radiation. The ultraviolet flux primarily responsible for the ionization of the E region has wavelengths between 900 and 1216 angstroms. This radiation is produced primarily by emission from excited hydrogen and carbon atoms in the solar atmosphere. The x-ray flux occurs in the wavelengths between 10 and 100 angstroms.[1]

Although diatomic oxygen comprises only 6 percent of the atmosphere at E-region heights, it furnishes about half of the free electrons. The other half of the free electrons are furnished by ions of nitric oxide $(NO+)$[2]. In addition, significant quantities of magnesium, sodium, iron and calcium ions have been found at E-region heights. The total electron density, N, in the daytime E-region has been found to vary as

$$N = 10^4 (180 + 1.44R)\chi$$

Where χ is the solar zenith angle and R is the smoothed sunspot number.

The intensity of solar radiation responsible for the E-region varies by a factor of about 2 between the years of low and high solar activity. The maximum electron density of the normal E-region is about 100,000 per cubic centimeter. The electron density of the nighttime E-region reaches a minimum of about 1000 per cubic centimeter which is comparable with the density of monatomic metallic ions. Since the characteristic time required for the recombination of the normal ions $(NO+$ and $O_2+)$ is on the order of 17 seconds and for monatomic ions is many weeks, it appears probable that the ionization of the nighttime E-region and nighttime sporadic-E is due to presence of meteoric metallic ions. [3, 4]

Fig. 1 is a time chart of E-layer muf for May 1969. This chart gives the maximum frequency usable for regular E-layer communications over 2000-km paths (1200 miles) and may be used in the same way as the F2 layer chart except the control point is only 600 miles away. Practically, however, the E-layer muf will frequently be much higher during summer due to the presence of sporadic-E (E_s).

sporadic-E

Sporadic-E is an ionospheric layer and propagation mode that involves reflection or scattering from the E-region of the ionosphere. Sporadic-E propagation is distinguished from regular refraction processes in the daytime E-layer in that higher than normal frequencies are reflected (often only partially) from regions localized in time and place. Sporadic-E is distinguished from meteor reflections in that it is prolonged over a period of several minutes to several hours.

Sporadic-E is observed both by vertical incidence ionosondes and by oblique sounders and communications circuits. Observations with an oblique sounder lead to a classification of nine different types of sporadic-E, eight of which can be found in the United States. The seasonal and diurnal characteristics of sporadic-E vary widely with latitude and even with longitude. In the auroral zone, sporadic-E is least likely during the hours of 0600 and 1500 local time; in temperate and equatorial zones Sporadic-E is most likely during those hours.[5] The boundary between the auroral and temperate zones, as far as Sporadic-E is concerned, is not well delineated.

My personal experience of over 15 years on 50 MHz suggests that the northern United States may have some characteristic of both zones. It is fairly well established that strong and widespread sporadic-E openings occur more frequently when the path midpoint is at southerly latitudes. In fact, sporadic-E occurs less frequently at north temperate latitudes than anywhere else. Texas, Florida, Southern California, and the Caribbean, the Gulf states and south-western states have an advantage over the northern states for multiple-hop 50-MHz and single-hop 144 MHz sporadic E.[6] However, for widespread occur-

rence of temperate-zone sporadic-E, the Far East is hard to beat. A very persistent type of sporadic-E also occurs over the geomagnetic equator.

A sporadic-E "cloud" is seldom a perfect reflector at frequencies of 50 MHz and above. Propagation losses of 20 to 60 dB over free space are common at vhf. One contributing factor is that the cloud may be small, i.e., less than a Fresnel zone* in extent. The primary factor, however, is that sporadic-E is frequently a poor reflector—most of the signal is transmitted through the "layer".

Present-day theories suggest that most temperate-zone sporadic E takes the form of a thin reflecting stratum, perhaps less than a mile or so thick, found most frequently at heights between 60 and 75 miles. The horizontal extent of a sporadic-E cloud may range from tens of miles to hundreds, and it may or may not move. It it does, it is generally from east to west at velocities of around 100 mph.[7]

Our prime interest in sporadic E is its ability to reflect signals at much higher frequencies than are reflected by normal ionospheric layers. Although sporadic E occurs at frequencies below 21 MHz, it is frequently not recognized as such. At 21 and 28 MHz, propagation frequently occurs by mixed sporadic-E and F2-layer modes. At 50 MHz there is little difficulty in recognizing sporadic E since it is usually the only mode propagating at the time.

May, June and July are the months of peak occurrence of sporadic E in temperate latitudes in the northern hemisphere. A minor peak frequently occurs in December and January.

A large patch of sporadic E as viewed by a vertical incidence sounder may evolve from a small bulge in the virtual-height-vs-frequency trace at the top of the E-region, descending and strengthening to form a thin layer near a height of 67 miles. Occasionally sporadic E is seen at heights as low as 55 miles, and more than one layer may be present at one time.

Only recently with the advent of sounding rockets and guns has appreciable progress been made in direct measurement of the conditions surrounding a sporadic-E cloud. Ionospheric winds are measured by observation of the drift of chemi-luminescent trails;[8] chemical composition of the positive ions is measured with a mass spectrometer.[9] The

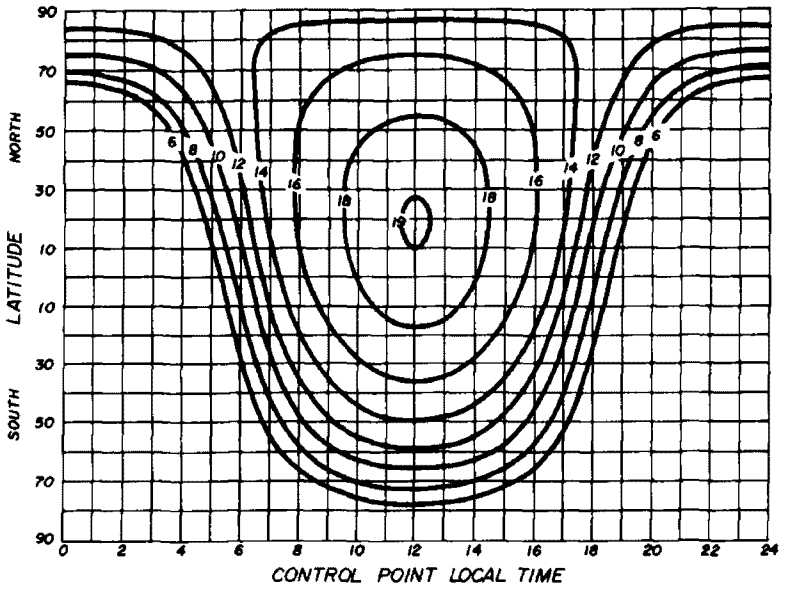


fig. 1. Time chart of median E-layer for May 1969 predicted for the regular daytime E-layer.

results of these measurements suggest that temperate-zone sporadic E results from an accumulation of ions at a preferred height of convergence of the vertical component of the neutral wind.[3]

sporadic-E last year

The year of 1968 was exceptional for the reported occurrence of sporadic E in the 144-MHz band. Openings were reported on May 29 and June 10, but the most widely observed were the openings of June 20 and 21, starting about 2230 gmt both nights.[10] The number of observations was sufficient for me to run a crude statistical analysis of the data. All contacts and "heard" reports listed in QST where both station's locations could be ascertained were computer analyzed to determine the path length and the coordinates of the path center.

Most of the reports were for paths between 900 and 1400 miles long (see **fig. 2**). A scatter plot of longitude and latitude of the path centers is shown in **fig. 3** with a rough outline of the state boundaries. Most of the path midpoints are clustered about the probable location of the sporadic-E cloud at the time of reception. Notable deviations are the contacts between K1FKW and K5GKR, and W3TFA and W4IID. Reports over 1400 miles and under 800 miles were:

W1AZK-W5MCC	1415 miles
K2BWR-W5GVE	1454 miles
W5MCC-W2YQI	1410 miles
W5MCC-W1YDF	1403 miles
W1AZK-W4LSU	749 miles
K2BWR-K4TAG	754 miles
W3TFA-K4TAG	633 miles
W4ISS-K2CBA	793 miles

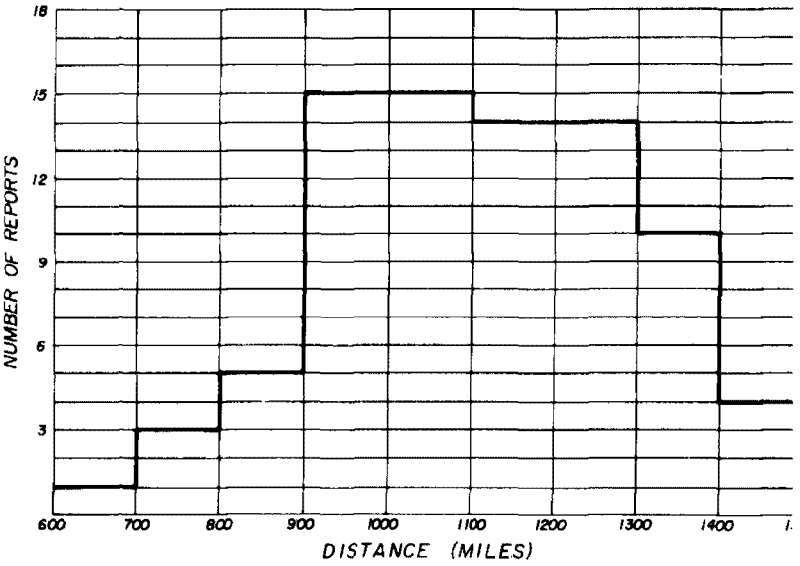


fig. 2. Distribution of two-meter reports according to distance for the sporadic-E openings of June 20 and 21, 1968.

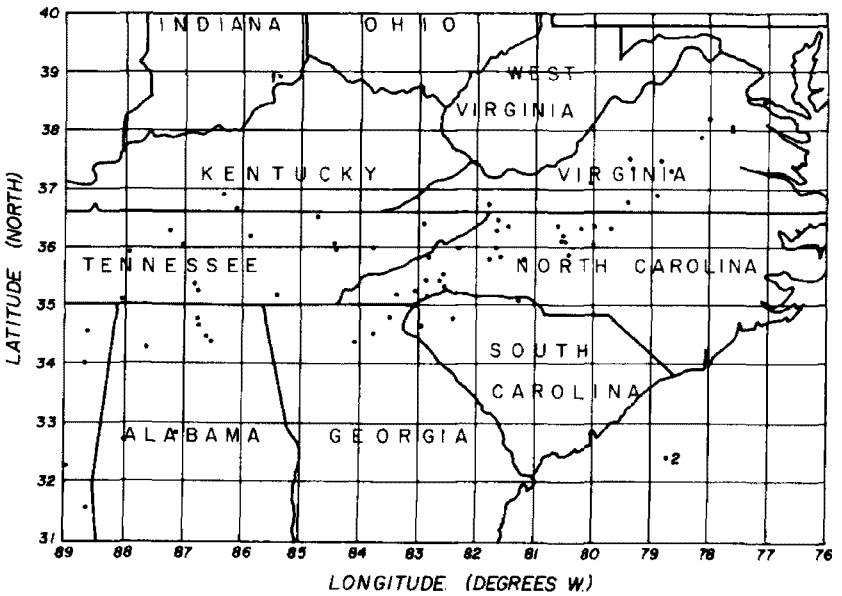


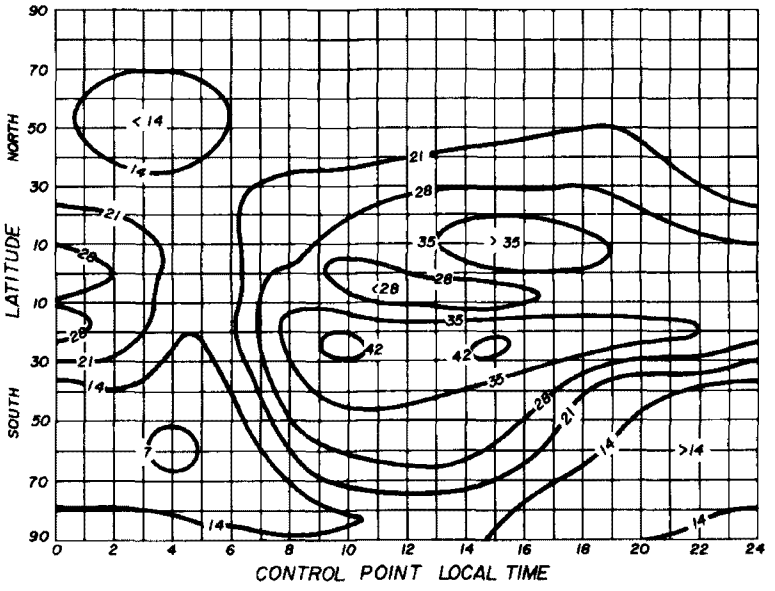
fig. 3. Midpoints of the two-meter paths for the June 20-21 openings. It is thought that the sporadic-E clouds originated somewhere east of Washington, D. C., and drifted to the Southwest. The points marked 1 and 2 indicate off-path and off-center reflections and are identified in the text.

Some of the exceptional distances or path midpoints may be due to errors in callbook addresses. Although I do not have the times of these reports, I would guess that the sporadic-E cloud was first detected just south and west of Washington, D.C., and that it spread and moved westward with time. Some of the spreading to the west may be due to different tracks on the two days.

of these charts have been covered in the column in the March issue.

80 meters: Daytime maximum ranges of 300 miles and nighttime maximum ranges of about 2000 miles may be realized in areas with low noise levels; but about half these distances will be more likely. May is a good month for local nets. Some DX may be

fig. 4. Time chart of median muf for May 1969 from ITS predictions centered on a longitude of 90° W.



It is very tempting to speculate that these two openings were more than a natural occurrence. Artificial electron clouds have been produced at Holloman Air Force Base, New Mexico,[11, 12, 13] Egland Air Force Base, Florida and Wallops Island, Virginia. However, a thorough check of operations at Wallops Island indicates no activity on these dates; thus sporadic-E events were most likely of natural origin.

propagation summary for may

Summer has arrived in the northern-hemisphere ionosphere, resulting in decreased daytime muf's, increased nighttime muf's and increased noise levels. The F2-layer muf's vs time of day and latitude are shown in **fig. 4.** May is one month where the differences between northern and southern states on 15 and 10 meters is quite evident. The maximum range charts are shown in **figs. 5** to **9.** The use

worked on quiet nights into the southern hemisphere, where wintertime conditions prevail, but frequent static crashes are likely to dampen your enthusiasm.

40 meters: Continuous daylight over the North Pole may attenuate the European and Asian broadcasters somewhat. Good nighttime propagation is expected to the Southern Hemisphere (Africa, South America and Australia). Midday maximum ranges may be as short as 600 miles.

20 meters: Twenty should remain open all night. Best DX will occur near sunrise (NW) and during the evening hours (NE). Remember that twenty may be closed to southern temperate latitudes during hours of darkness (theirs). During midday, however, you may have trouble working across the continent.

15 meters: Fifteen will be the optimum band

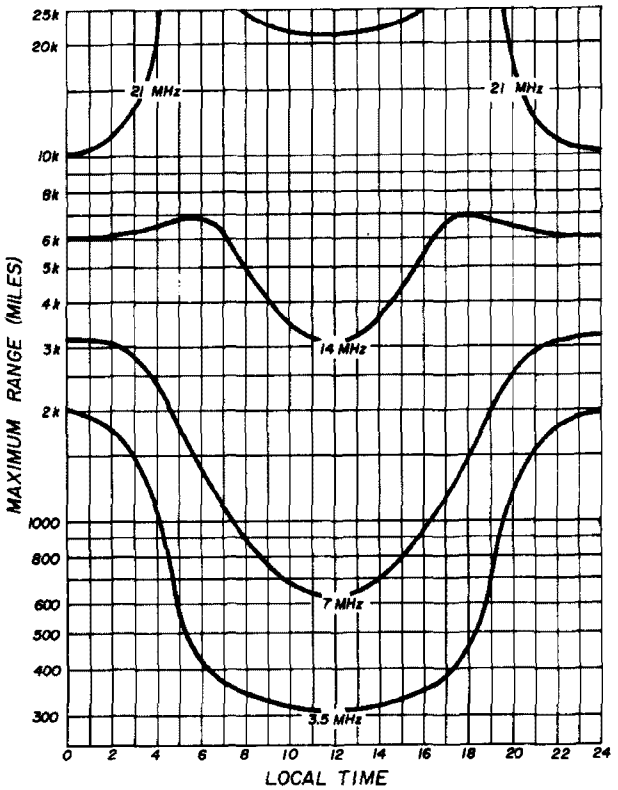


fig. 5. Maximum range to the north from 38° N. latitude as limited by absorption, atmospheric noise and system parameters (100 watts cw to typical antennas).

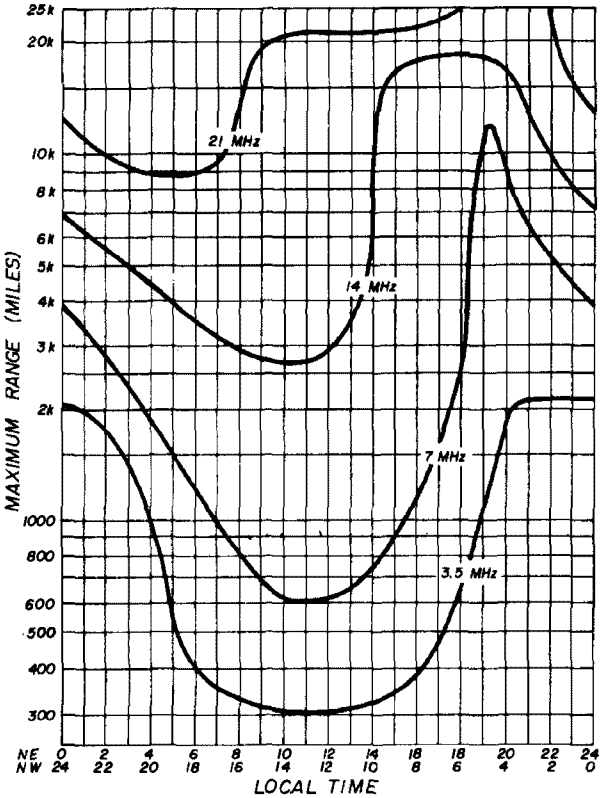


fig. 6. Maximum range to the northeast (top time scale) and to the northwest (bottom time scale) from 38° N. latitude.

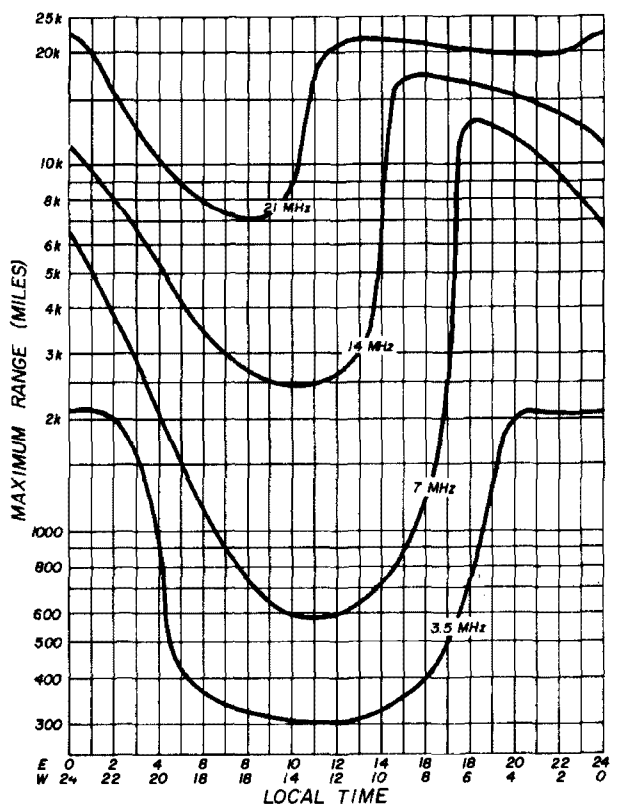


fig. 7. Maximum range to the east (top time scale) and to the west (bottom time scale) from 38° latitude.

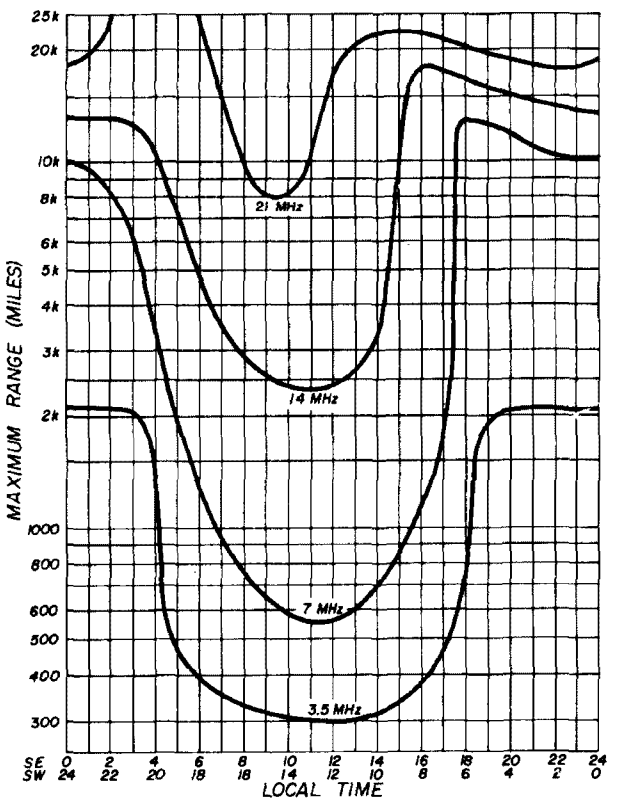


fig. 8. Maximum range to the southeast (top time scale) and to the southwest (bottom time scale) from 38° N. latitude.

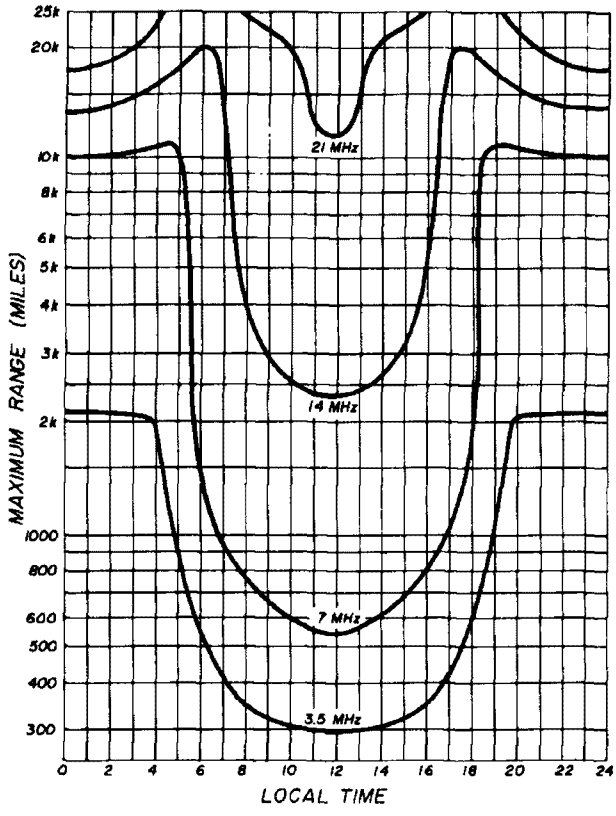


fig. 9. Maximum range to the south from 38° N. latitude.

for daytime DX, but may not open to other than southerly directions until midmorning. The time of the highest muf's to the northeast may be as late as 2 pm and to the northwest may be as late as 8 pm. Fifteen may be open to the southwest as late as midnight.

10 meters: Many will have given up ten meters since openings between east and west coasts may have become much less reliable than during the Spring. However, muf's are way up in the daylit southern hemisphere and if you want to work Antarctica on ten, this is a good month. Sporadic E will be in evidence by one or more hops.

6 meters: May marks the end of the transequatorial (TE) season although ZK1AA reports TE reception of Hawaiian tv signals throughout the summer months. May also marks the beginning of the sporadic-E season, which will likely start at southerly latitudes early in the month and work north as the month progresses.

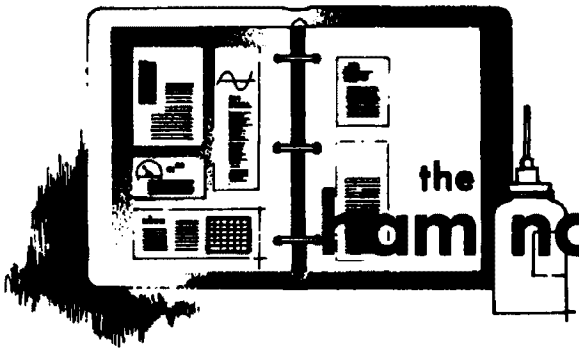
2 meters: Two meter operators should again be on the lookout for sporadic E, especially during the daylight and early evening hours, at southerly latitudes, and towards the end of the month. Tropospheric openings will be occurring farther north, and by the time this appears in print, KH6EEM and I should be running regular schedules trying to rebreak the California-Hawaii tropo path.

meteor showers: Include the Aquarids, May 1 to 6 (0300-1200); the Cetids, May 19 to 21 (0530-1430); the Herculids, May 11 to 24 (1800-0630); and the Pegasids, May 30 (2300-1200).

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ham radio



bandswitching fet converter postscript

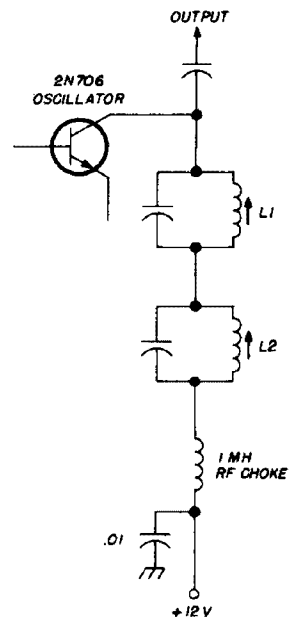
Since I recently completed the development of the "second generation" bandswitching fet converter,[1] I felt that several design changes included in it would be of interest to those who built the initial version or are working along similar lines.

Being rather elated at the time when Motorola solved my instability problems with a new super mosfet, it was only natural when contemplating a new design to give them a call on the off chance that they had come up with bigger and better things since my first attempts—and they had. The rf amplifier in the new version sports a mosfet with feedback capacity rated at less than 0.03 pF while retaining the high gain and low noise figure of the device I used in the previous converter.

In addition, the new fet is a dual-gate affair, allowing agc voltage to be applied to the second gate. The specs state that —3 volts will cut off the device. These new transistors can be used in the original converter but if you don't want to apply agc, the second gate should be connected to the positive supply (not greater than 24 volts).

The second design change occurred in the output circuit of the oscillator and was initiated after a discussion with the crystal manufacturer. He informed me that most high-frequency crystals above 20 MHz were overtone types and a selective circuit in the oscillator was necessary to pick out the correct overtones. It's possible that images could appear if this is not done, and if anybody experienced images on ten meters with the previous circuit, a tuned circuit will probably cure the problem.

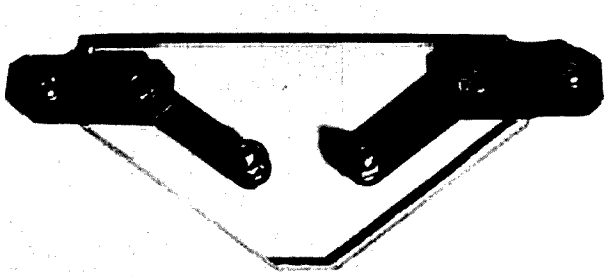
I had trouble with twenty-meter images in the new converter when listening on fifteen, although mathematically they should have been outside the range of the i-f strip being used to test the converter. I added a similar filter in the oscillator circuit for fifteen meters and cured the problem.



As seen in **fig. 1** the oscillator output circuit now appears as two parallel-tuned circuits in series with an rf choke. The parallel circuits have a very high impedance at their respective resonant frequencies, but a minimal impedance at frequencies removed from resonance. They are adjusted for resonance at the crystal frequencies used on the ten- and fifteen-meter bands. Except for these changes, the second generation converter is identical in all aspects except for size and the inclusion of a power supply.

Mike Goldstein, VE3GFN

dipole center insulator



If you need a dipole center insulator in a pinch and can't get down to your local electronics emporium, try the gadget shown in the photo. Just cut out a triangle from 1/4-inch plexiglass or other strong plastic, install some copper or brass strips with brass nuts and bolts, hook up your antenna wire and feedline and you're in business.

Ted Woolner, WA1ABP

using integrated circuits

There are many amateur projects showing up with integrated circuits, but IC's in the dual-inline package (DIP) have two rows of pins spaced 0.1 inch apart. This tight spacing makes casual breadboarding difficult at best, even if you use printed circuits. Here are two simple methods you can use to make the most out of IC's.

Use standard Vector number 169P59/032 perforated board. This board uses a hole arrangement that will fit either the fourteen- or sixteen-lead dual-inline package. Although it is fairly expensive, you can plug your IC in, solder wires to each pin and bring these out to flea-clip. This approach is a natural for casual experiments.

The second method is to turn the IC upside down and secure it to the board with a small dab of glue (be sure to note the alignment dot on top of the IC before turning it over). This method lends itself to circuits that are going to be used for awhile. When soldering to the IC pins, use a small well tinned iron and gingerly solder your connections. Normally no heat sink is required if you are careful and treat the device as you would a silicon transistor.

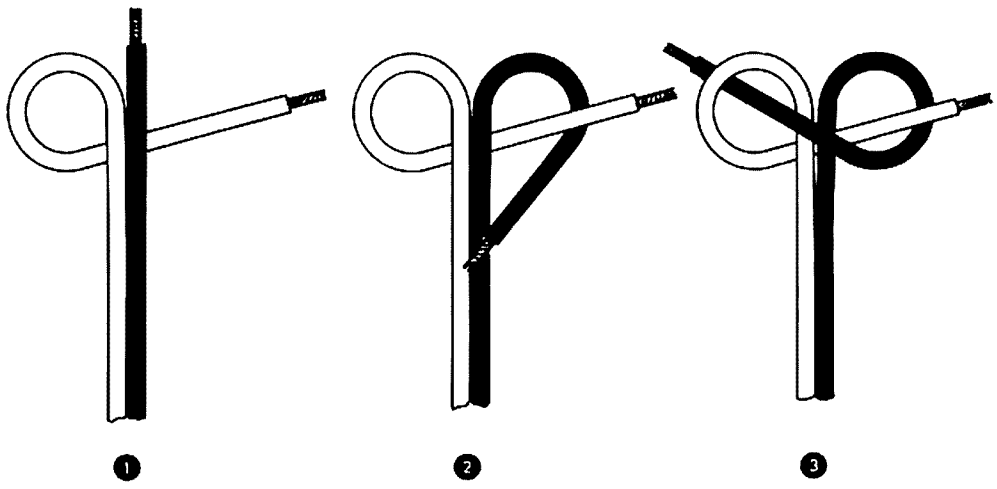
Elliott Kanter, W9KXJ

underwriter's knot

I'd bet most hams are pretty adept at replacing ac power plugs and installing new line cords, but how many know how to tie the underwriter's knot? Most hams use a simple over-hand knot when terminating a line cord in a plug or lamp socket, but this is not only incorrect, it's unsafe. As you can see from the three steps shown in **fig. 1**, the underwriter's knot is easy to tie when you know how. Use it next time instead of that half-knot you've been using all these years—it's a lot safer.

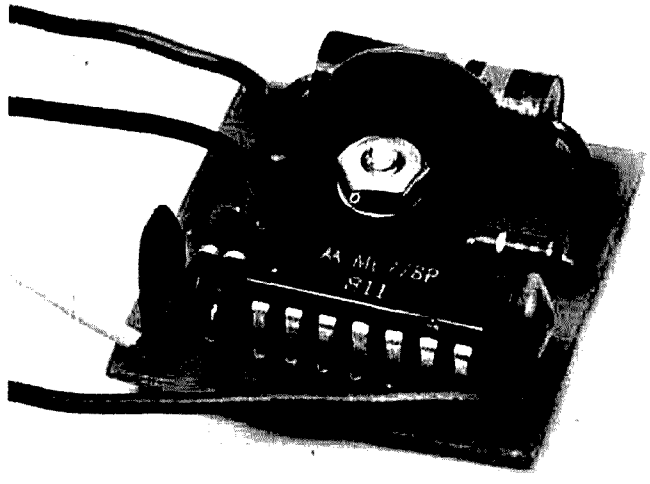
Jim Fisk, W1DTY

fig. 2. Underwriter's knot.



paxitronix frequency calibrator

Since the new frequency regulations went into effect, a number of magazine articles have described ways of getting 25 kHz check points from a 100 kHz calibrator. Although tubes or transistors could be used for this task, integrated circuits can do the job with more accuracy and at less cost. The Paxitronix IC-3 uses a dual JK flip-flop IC in a divide-by-four circuit that will convert a 1.5 volt rms sine wave into a square wave at one quarter the input frequency. The maximum toggle frequency of the MC778 flip-flop is in the neighborhood of 8 MHz so it works well with the 100-kHz crystal calibrators that are found in most modern communications receivers.



The Paxitronix IC-3 frequency divider consists of a transistor amplifier stage and integrated circuit, all on a miniature printed-circuit board about 1.3 inches square. Also mounted on the board is a 2-watt dropping resistor—its value depending on the dc voltage that is available to run the IC-3. Total current drain is very low so it's no problem to tack the unit on to the existing dc supply in your receiver. Although the IC-3 can be used with dc power supplies up to 300 volts, lower voltages are recommended by the manufacturer—down to about 3 volts dc. Most units will work satisfactorily down to about 1.5 Vdc.

The lower voltages are recommended because of the heat generated by the large dropping resistor that is needed with typical receiver power supplies. If you don't have a source of low-voltage dc available in your receiver, you might consider rectifying the 6.3 Vac filament supply for the job—all it requires is one diode, one resistor and one small dual-section electrolytic capacitor; Paxitronix gives you a circuit on the data sheet that comes with the IC-3.

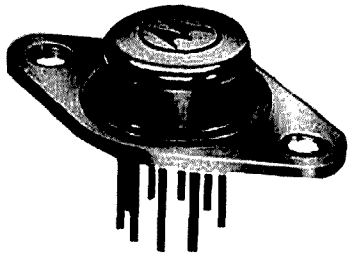
Although the Paxitronix IC-3 is only rated to 30 MHz by the manufacturer, I could detect 25 kHz markers to nearly 60 MHz, so the unit would serve well on six meters. Usable markers are available with 3 mA current drain, but much stronger markers result with 5 to 10 mA of current.

The IC-3 is simple to install. Only one mounting screw is required; four wires take care of the electronics—input, output, and two for the power supply. If you've been meaning to build a frequency divider but haven't gotten around to it, the IC-3 is a natural. $7.25 postpaid from Paxitronix, Inc., P. O. Box 1038(D), Boulder, Colorado, 80302.

Jim Fisk, W1DTY

new products

ic voltage regulators



A new integrated-circuit voltage regulator from Motorola features 0.002% regulation and output current up to 500 mA in a single package. The new regulator, the MC1560, is available in full-temperature-range tight-specification versions as well as a relaxed-spec unit, the MC1460, that sells for as little as $5.25.

One of the big features of the MC1560 is its very low output impedance—typically 20 milliohms (0.020 ohm)—and this varies only a few milliohms over the output voltage range of 2.5 to 17 volts; maximum input voltage is 20 volts. Temperature compensation is provided by a resistive divider on the monolithic chip that balances the positive temperature coefficient of a zener diode against the negative temperature coefficient of forward-biased diodes.

Another feature of the MC1560 is the shut-down control. If a control voltage is applied to the shut-down terminal, both the load and regulator bias current are turned off; this can be used for remote on-off switching, for squelch control in communications equipment or to protect the regulator during sustained short circuits.

The data sheet for the MC1560/MC1460 offers complete information to the designer: it not only lists ratings and electrical characteristics with extensive notes on measurements, but also has a full page of operating data. The MC1460R is in a new 9-pin version of the TO-66 power transistor case that will dissipate 10 watts at 65°C and regulate a 500-mA load. A single external power transistor can increase this to more than 10 amperes. The MC1460G, packaged in a 10-pin TO-5 case, will dissipate 1.8 watts at 25°C and regulate a 200-mA load. The MC1460G is $5.25 and the MC1460R, $6.75 in small quantities.

The Motorola MC1461 is a voltage regulator IC with all the features of the MC1460 but with an increased maximum input voltage limit—up to 35 Vdc. Regulated output may be set from 3.5 volts to slightly over 30 volts, depending on the external components. Packaging and dissipation ratings are the same as the MC1460. The MC1461G sells for $6.75 and the MC1461R, $8.25 in small quantities. For more information on the MC1460 and MC1461 voltage-regulator IC's, write to Technical Information Center, Motorola Semiconductor Products, Inc., P. O. Box 20924, Phoenix, Arizona 85036.

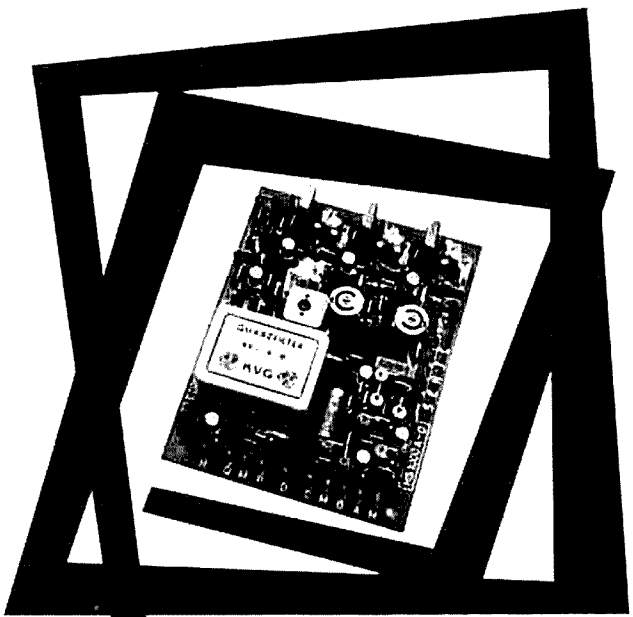
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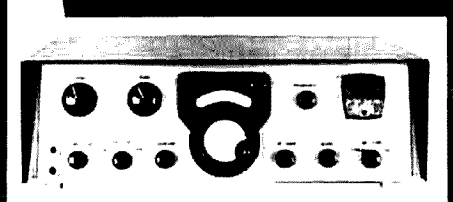
focus
on
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technology . . .

ham radio magazine

JUNE 1969



SOLID-STATE single-band SSB transceiver



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ham radio magazine

june 1969

volume 2, number 6

staff

editor
James R. Fisk, W1DTY

roving editor
Forest H. Belt

vhf editor
Nicholas D. Skeer, K1PSR

associate editors
A. Norman Into, Jr., W1CCZ
Alfred Wilson, W6NIF
James A. Harvey, WA6IAK

art director
Jean Frey

publisher
T. H. Tenney, Jr. W1NLB

offices

Greenville, New Hampshire 03048
Telephone: 603-878-1441

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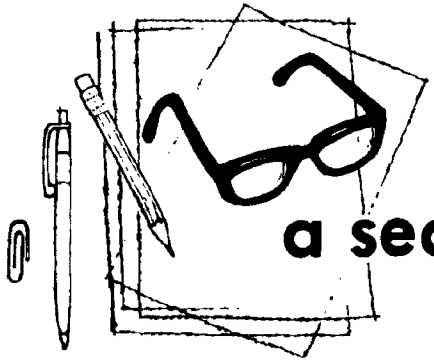
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a second look

by jim fisk

It seems that more and more stations are getting ready for serious moonbounce work. And by the end of the summer the number will probably double. Stations capable of eme work were pretty scarce three or four years ago—only the most serious and perservering workers ever made the grade. But techniques have been improved, and probably more important, equipment is available so just about anyone can hear his own two-meter echos from the moon.

In the days right after World War II, when the first radio signals were bounced off the moon by Army experimenters at Fort Monmouth, New Jersey, it wasn't nearly so simple: 8 kW input on 111 MHz, a tremendous billboard antenna with 64 phased dipoles and an incredibly complex receiver with a 50-Hz passband. Even then the moon echos were weak, and success unpredictable. (To improve reliability, the Army eventually increased transmitter power to 100 kW.)

With this kind of a background, the prospects of amateur communications via the moon were pretty remote, but W3KGP and W4AO launched Project Moonbeam in the late 1940's with a goal of bouncing two-meter signals off the lunar surface. All indications were that a successful effort would require the full amateur power limit, antenna gain of at least 20 dB and receiver performance that was practically unattainable. But in July, 1950, came something like an echo, faint and indefinite, but it sounded like the real thing and it was captured on a wire recorder. Test after test followed, with failure after failure, but finally, 2½ years later, and many equipment changes in between, they managed to record a whole series of echos off the moon.

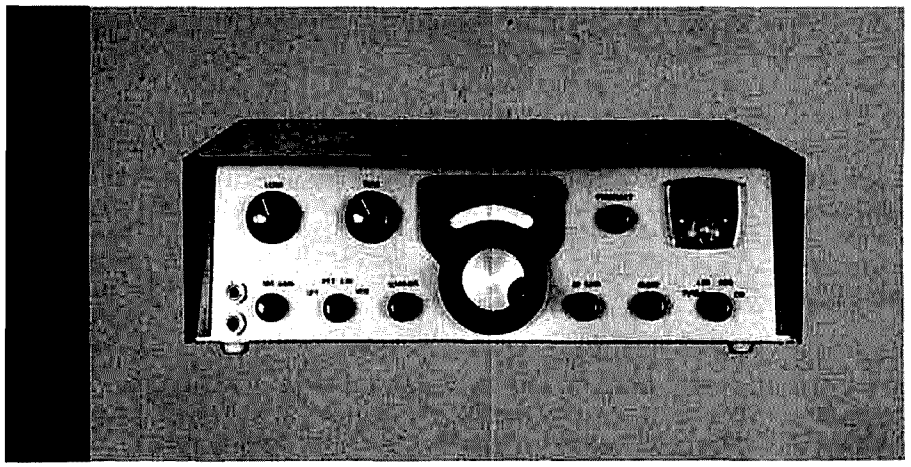
After this initial amateur success, progress was slow. Many amateurs tried, but few succeeded. There was a flurry of activity as amateurs built bigger and bigger two-meter arrays and high powered amplifiers—but successes were limited and there were no two-way eme communications.

Until practical parametric amplifiers became available, moonbounce activity was confined to two meters. Even with a low-noise front end, moonbounce attempts on 432 MHz were impractical because of the power limitation, so the next logical step was 1296 MHz. After a lot of sweat and tears, W1BU came on the air in early 1960 with a 1296-MHz station that could bounce signals off the moon with some degree of reliability. W1FZJ extended the challenge to vhf enthusiasts; Hank Brown, W6HB, picked it up, and shortly thereafter the first two-way amateur contact via the moon's surface was history.

Since then progress has been slow but steady. Nearly all the vhf bands up to 1296 MHz have some sort of moonbounce activity, but at the moment two meters is the most popular. Nor is activity confined to the United States; successful moonbounce stations are located in Australia, Finland, France, Greece, New Zealand and Sweden.

Moonbounce is still a very sophisticated method of vhf communications, but on two meters at least, it is within the grasp of any serious worker. Moonbouncers K6MYC and W6DNG live on city-sized lots, so space limitations are apparently no problem. KØMQS uses all commercial equipment except for the antenna, so equipment is not a problem. The only other major problem is perservance—but nevertheless, more and more stations will be showing up on the low end of two with their antennas pointed toward the moon. And in the not to distant future, working all 50 states on two meters may even be a reality.

Jim Fisk, W1DTY
Editor



homebrew single-band ssb transceiver

Complete construction details for the HR ssb transceiver—an easy-to-build modern design for 75 meters incorporating integrated circuits, transistors and a vacuum-tube final

Jim Fisk, W1DTY, Box 25, Rindge, New Hampshire 03461

Although solid-state has been with us for a long time, there have been very few reproducible solid-state transceiver designs in the amateur literature. Most of the homebrew ssb transceivers that you hear on the bands—although they are few and far between—are tube types. In fact, the only station I have worked using a homebrew solid-state ssb transceiver was UF6ACR in the U.S.S.R.

Evidently solid-state is new enough that few amateurs are willing to tackle a project as large as a transceiver. The big hangup, of course, is the sideband generator; amateurs are willing to build microphone preamps, electronic keyers and rf power amplifiers, but when it comes to balanced modulators and crystal filters, they stop.

When Spectrum International* started advertising miniature solid-state sideband exciters several months ago, I was really intrigued. A quick note brought more detailed literature. The ssb exciter, made in West Germany by DJ3CI, includes a three-stage microphone preamp, crystal-controlled oscillators, balanced ring modulator and a 9-MHz crystal filter. All the hard work is done—here is a miniature solid-state source of 9-MHz ssb with 55 dB carrier suppression and 45 dB suppression of the unwanted sideband. In ad-

dition to the selectable sideband capabilities of the module, it has provisions for a-m, fm and CW.

I was particularly interested in the KVG crystal filter used in the transceiver module. This 8-pole crystal filter is very popular in Europe. Its excellent passband characteristic has a shape factor of 1.66:1 from 6 to 60 dB down; minimum stop-band rejection is greater than 80 dB.

duced to a balanced mixer stage along with the output from the vfo crystal-oscillator circuits. For operation on the 3.5 and 14 MHz bands, no crystal oscillator or mixer stage is required; the output from the 5.0 to 5.5 MHz vfo is injected directly into the balanced mixer.

The output of the balanced mixer is amplified by several broadband amplifier stages. The transistors in these stages are run in class

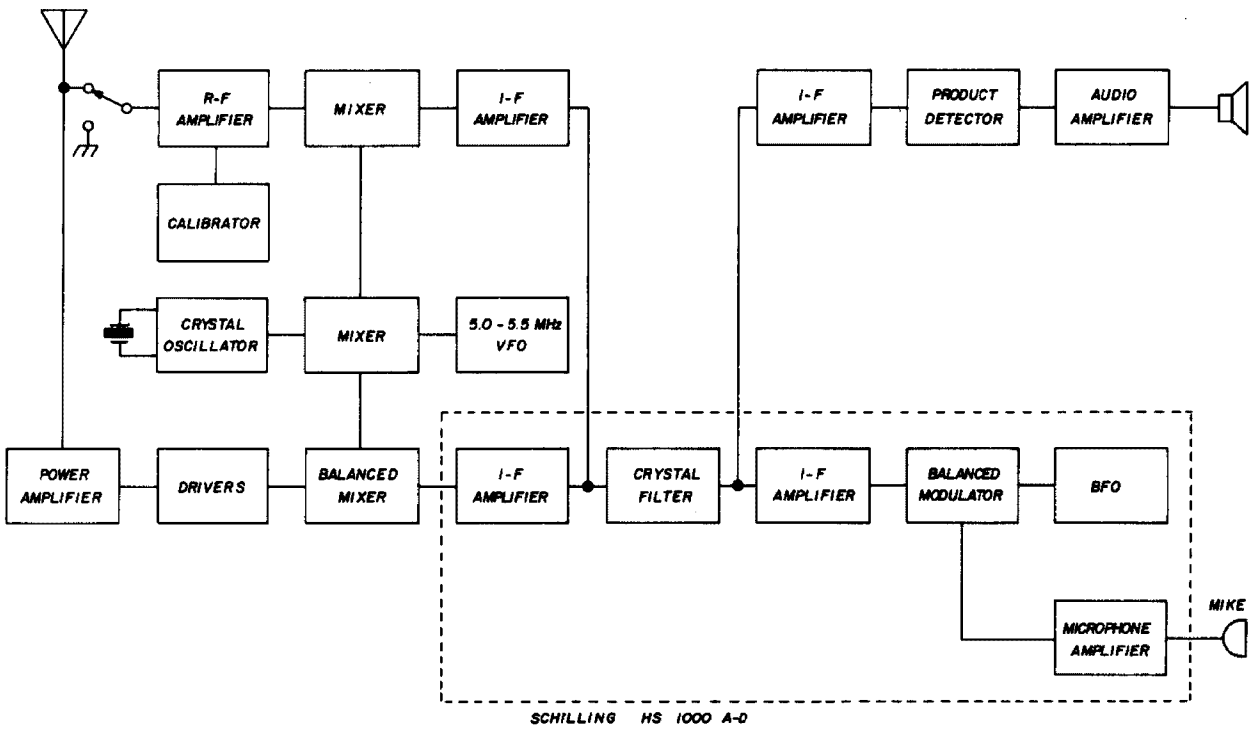


fig. 1. Block diagram of the single-band transceiver.

design

To keep things as simple as possible, the initial design was for a single-band unit. This is in keeping with most mobile and portable operation and minimizes the mechanical problems of bandswitching. Also, with the single-band approach, the same basic design philosophy can be applied to the vhf bands with very few changes.

A block diagram of the complete transceiver is shown in **fig. 1.** The imported ssb exciter module is enclosed in a dotted box. The 9-MHz output of the module is intro-

* Spectrum International, Box 87, Toppsfield, Massachusetts 01983

B for minimum distortion and maximum efficiency. Gain is lower in this configuration, but linearity is good, and no external tuning is required. If you are willing to put up with an external control for peaking up these driver stages, you could probably delete one of the broadband amplifiers. These transistors are inexpensive, though, and for my money the operating convenience of minimum tuning is worth the additional stage.

The vfo uses the basic fet Seiler oscillator described previously[1] followed by a class-A buffer stage and emitter follower. For operation on 80 or 20, this is all that is necessary.

The final power amplifier provides about

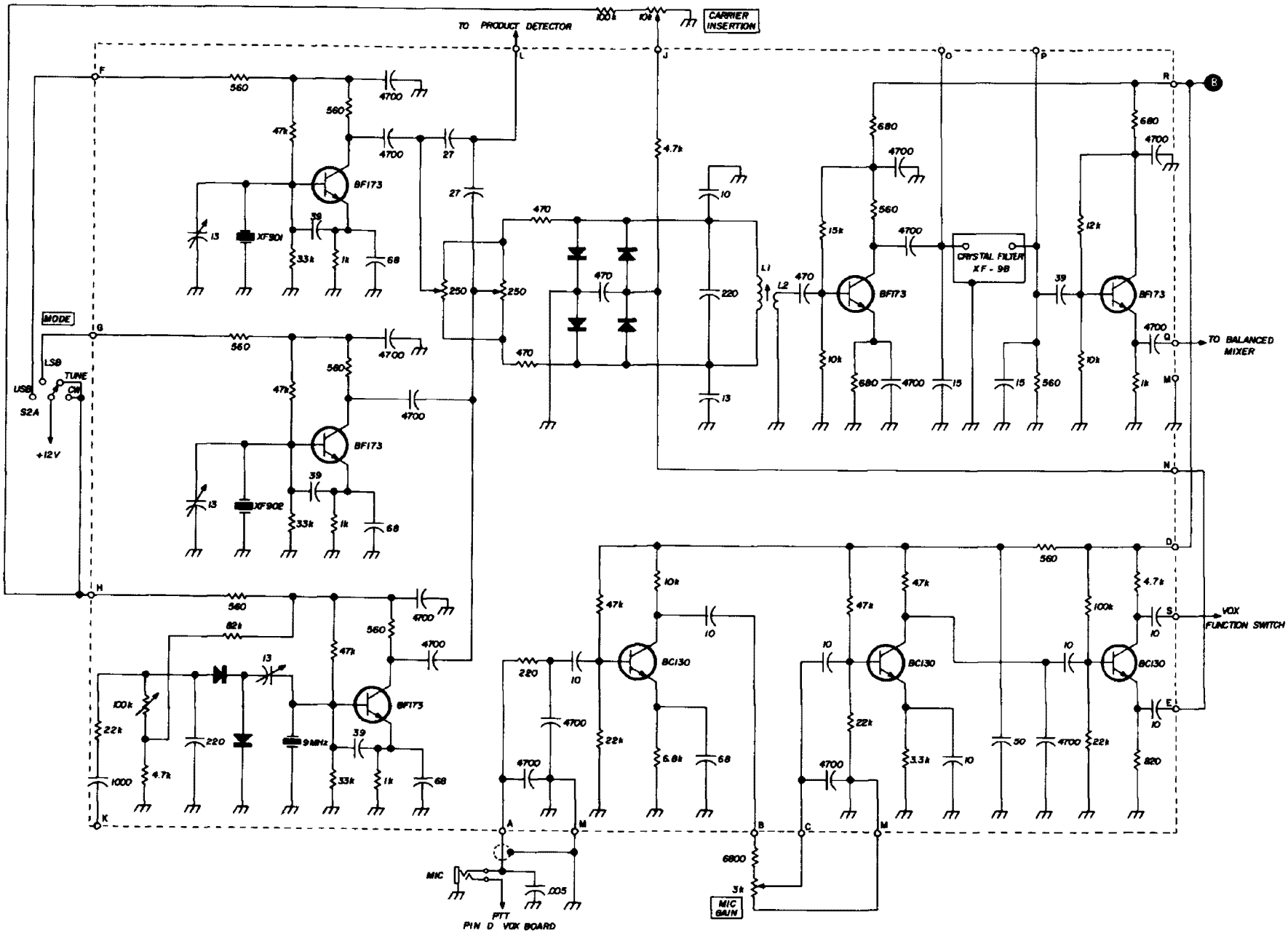


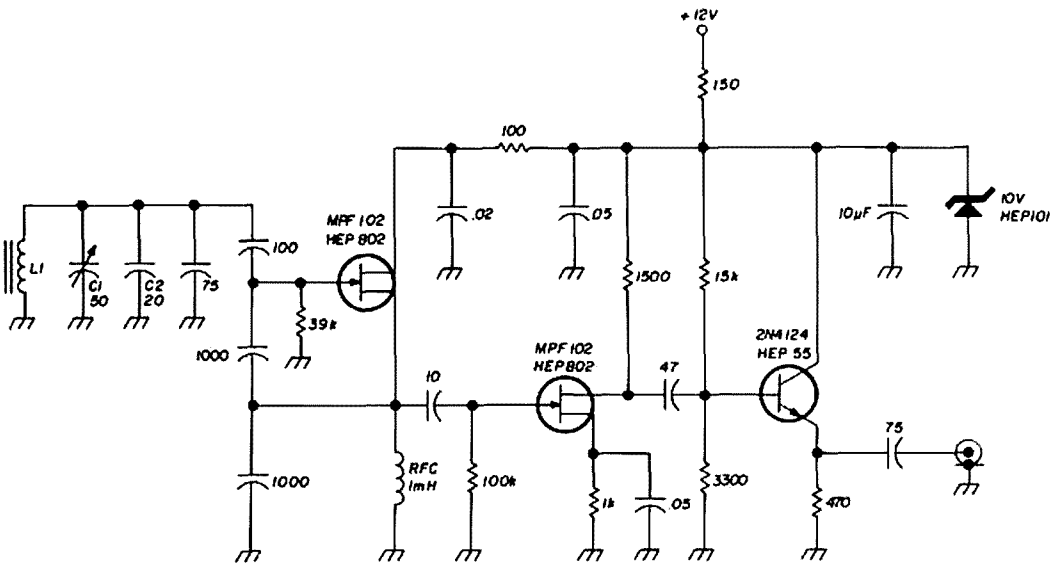
fig. 2. Schematic diagram of the Schilling HS 1000D 9-MHz transceiver module. Components outside the dotted line must be added by the builder.

50 watts PEP output. There are several transistors on the market designed for ssb service that will provide this much output, but I felt that for the present, a vacuum tube was much more feasible. The physical layout of the transceiver is such that when the cost of high-power rf transistors drops within reason, the vacuum-tube power-amplifier assembly can be replaced with an all solid-state unit.

The receiver lineup is pretty conventional —mosfet rf stage followed by a mixer into the 9-MHz crystal filter. Output from the filter is amplified in an integrated-circuit i-f amplifier and fed to a product detector and the audio stages; agc voltage is derived from the audio section and provides excellent dynamic range.

transmitter section

After you receive the exciter module, you'll probably be pretty anxious to get it hooked up and working. I know I was. You can put it on the air in a matter of minutes if you have a 12-volt power supply and a handful of



C1 50-pF variable (Millen 21050MK)*

C2 20-pF variable, printed-circuit type (Jackson Bros. 5440/PC/20)

L1 29 turns no. 24 enamelled on a toroid form (Amidon Associates T-50-2)

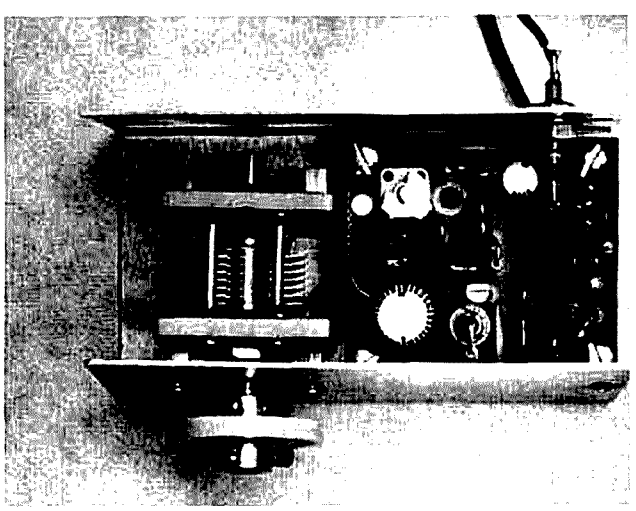
fig. 3. The Seiler vfo circuit tunes from 5.0 to 5.5 MHz. The 75-pF capacitor consists of a 33-pF, N750 temperature-compensating capacitor in parallel with a 42-pF silver mica.

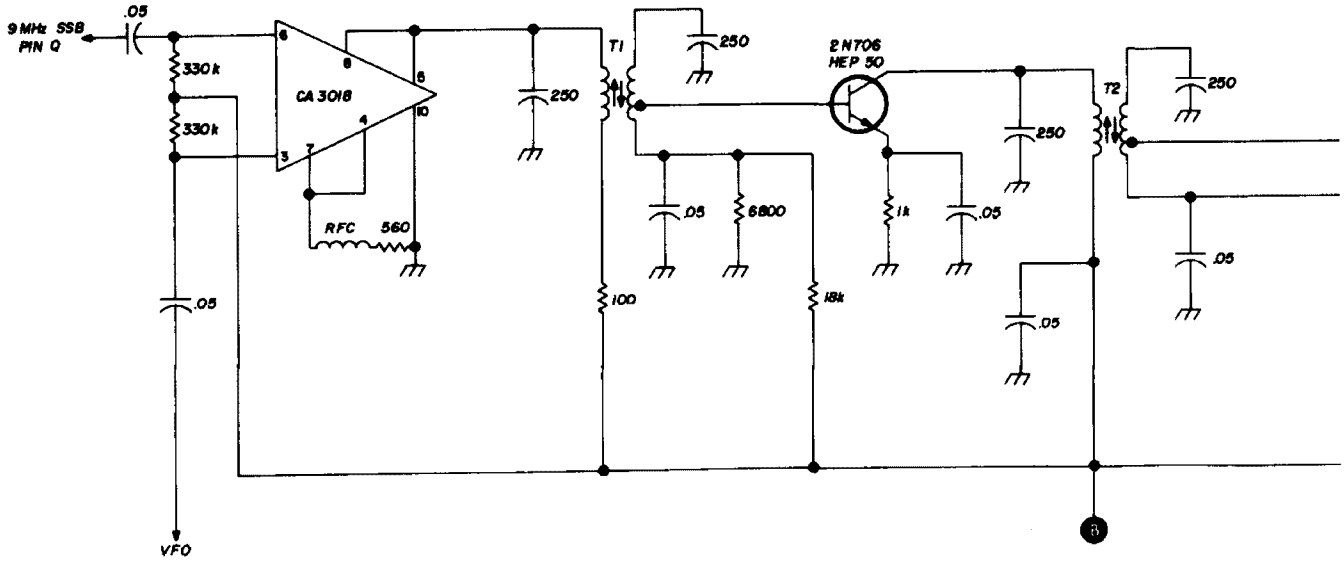
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* Millen parts are available from James Millen Manufacturing Company, Inc., 150 Exchange Street, Malden, Massachusetts 02148. Millen 21050MK double-bearing capacitor is priced at $2.91; the 39016 flexible coupling is $.75.

Jackson Brothers components are available from Arrow Electronics, Inc., 97 Chambers Street, New York, New York 10007; 5440/PC/20 capacitor is $2.50.

Complete vfo and buffer circuits are built on a small board mounted in a 2x2x4-inch minibox. Millen tuning capacitor and flexible coupling are on the left.





C3 20-pF (part of Miller 1460, see fig. 5)

C4 100-pF variable (Hammarlund HF-100)

C5 150-pF, 500 Vdc

C6 730 pF (parallel-connected dual-section 365-pF trf receiver capacitor)

C7 1000 pF mica

L7 30 μH (65 turns no. 30, scramble wound on a ¼" slug-tuned form, tapped at 20 turns)

L8 9.5 μH (32 turns no. 14 enamelled on a 2" toroid form—Amidon Associates T-200-2)

T1, T2 Primary is 33 turns no. 32 on ¼" slug-tuned form, tuned with 250 pF; secondary is 33 turns no. 32 on a slug-tuned form spaced ½" from primary, 8 turns from ground, tuned with 250 pF

T3

PS1 Parasitic suppressor, 8 turns no. 20 wound on body of 100-ohm, 1-watt carbon resistor

K1 spdt relay, 12 Vdc coil, mounted in power-amplifier compartment

fig. 4. Transmitting mixer, drivers and power amplifier.

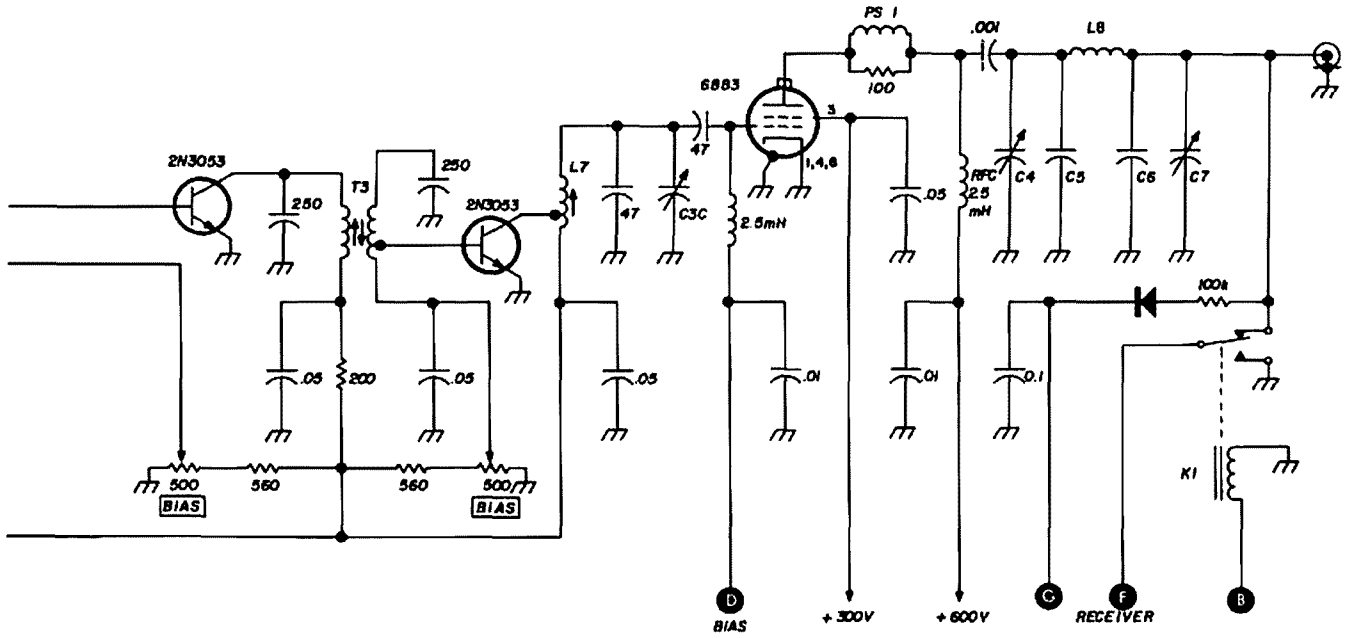
clip-leads, but it only takes a little while longer to do it right with a microphone gain control and a mode switch. If you bought the vox module, you can wire this in, too, since it only takes a few minutes to make the connections. The vox module is a good buy because it gives you built-in vox with anti-trip along with push-to-talk and a 4pdt relay to handle the transceiver switching functions.

The mode switch in **fig. 2** selects the applicable crystal oscillator for upper or lower sideband or CW. In the CW and tuneup positions, carrier insertion is adjusted by the 10 kilohm control. If you have a receiver that will tune to 9 MHz, you should be able to hear yourself on 9-MHz single sideband without even connecting an antenna to the receiver. If you don't have a general-coverage receiver, couple your grid dipper to the output—it's sufficient to pin the needle when the gdo is tuned to 9 MHz.

vfo

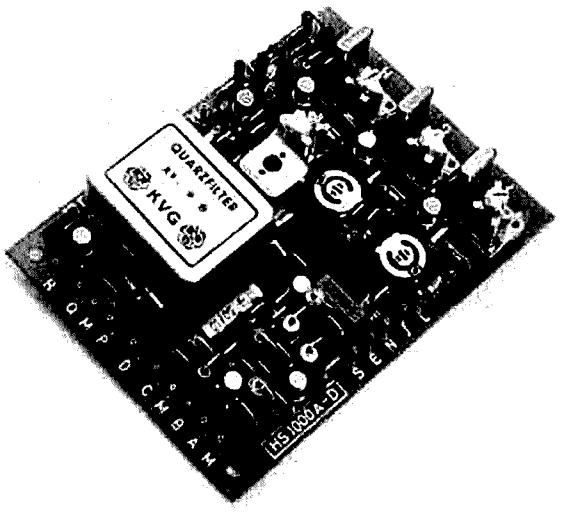
After you get the exciter module going, the next logical step is the 5.0- to 5.5-MHz variable-frequency oscillator. The vfo and buffer stages are mounted on a piece of printed circuit board and mounted in a 2x2x4-inch minibox. The heart of this very stable circuit is the Seiler oscillator; the output level varies less than 0.5 dB when the vfo is tuned through its range, and resetability is excellent.

The buffer stage uses an fet operated in class A to minimize loading on the vfo. The emitter-follower provides a low-impedance output for driving the miniature 50-ohm co-



axial cable to the balanced mixer; maximum output is 2.5 volts p-p. The 75-pF output coupling capacitor should be adjusted to give 1 volt p-p at the input to the balanced mixer.

Temperature compensation requirements are minimal because of the high stability of the variable oscillator circuitry. In this circuit, stability is primarily a function of the tuned-circuit components and is only slightly affected by the active device. A 33-pF N750 negative-temperature coefficient capacitor in the tank circuit should take care of any temperature-induced drift.



This 9-MHz ssb generator is the heart of the transceiver.

transmitting mixer

The transmitting mixer uses a low-cost integrated circuit in a balanced configuration. An earlier design used a junction fet with injection to the source and gate, but spurious outputs and carrier feedthrough were a problem. When I switched to the balanced circuit shown in fig. 4, these problems disappeared. With a properly operating balanced mixer circuit, little 5-MHz carrier and 9-MHz ssb appear across the output so problems with intermodulation distortion from this source are eliminated. Also, carrier feedthrough is a serious problem with bipolar transistors; before the balanced mixer circuit was used, the 5-MHz vfo signal fed through to the final.

Although an integrated circuit is not absolutely necessary in this spot, the transistors should be closely matched for best operation. If you have a matched pair of npn transistors with suitable frequency response they should work as well as the CA3018. I chose the IC because it was inexpensive and its transistors are practically identical since they are made on the same chip of silicon.

The bandpass coupler on the output of the balanced mixer is designed to pass the entire amateur band for which it was designed. This is most difficult on 80 meters where 500 kHz is a big percentage of the center frequency; on the higher bands it is somewhat easier to achieve.

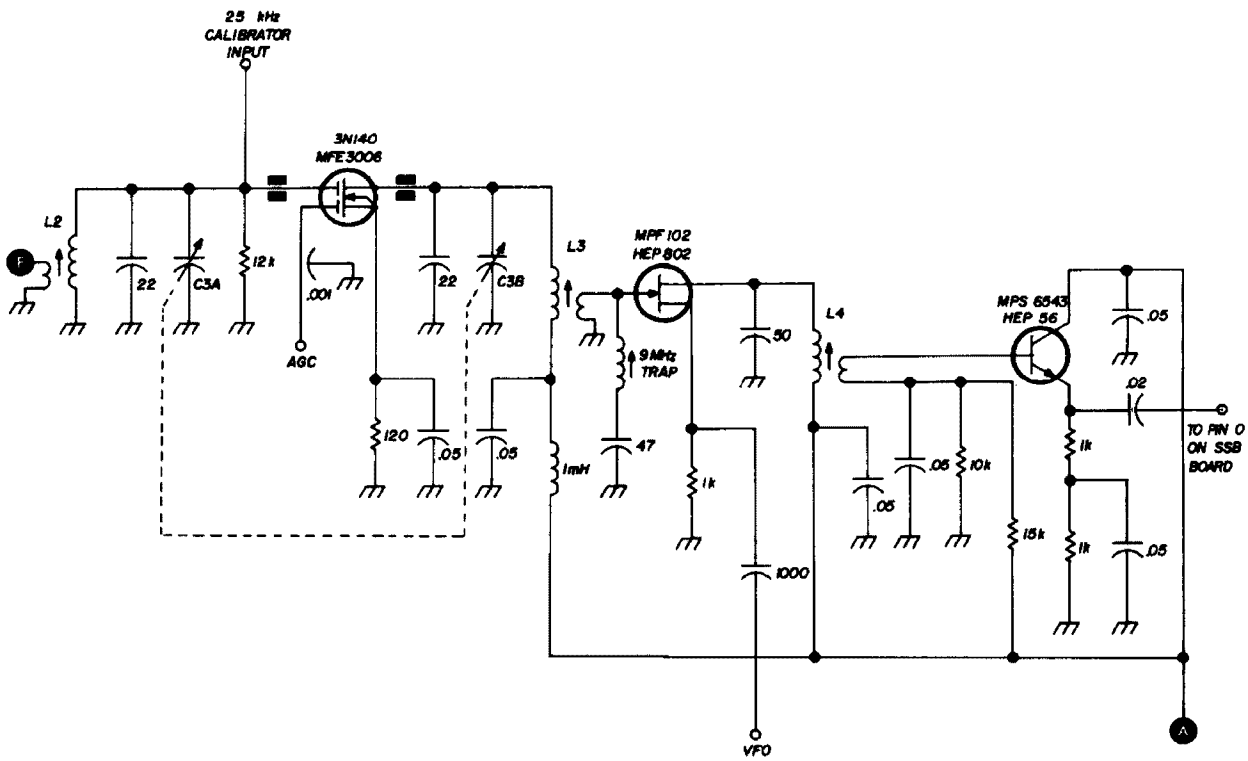
transmitter drivers

The transistor driver stages are operated in class B for maximum linearity and efficiency. This is accomplished by slightly forward biasing each driver stage. The bias supply must be very "stiff" so that bias remains constant with varying input signal; consequently, be careful when setting up stage bias. If the forward bias is set too high, the transistors will destroy themselves by thermal runaway. However, this is no problem if you put a milliammeter in the collector supply lead when setting the no-signal collector current.

The correct amount of no-signal bias current varies slightly from one transistor type to another, but will be in the range from 2 to 10 mA. If the forward bias is set too high, it will be immediately apparent if you have a meter in the circuit—with no input signal, the collector current will creep upward, slowly at first, and then alarmingly fast, until the transistor fries itself. However, if you're aware of the problem, you shouldn't lose any transistors. Furthermore, once the bias is set up it should require no further attention. Correct no-signal collector current for the 2N3053's is approximately 5 mA.

I used 2N3053's in the driver stages because they're inexpensive and I had some on hand. They also have frequency response and voltage ratings that are appealing. Power gain of typical units starts to fall off in circuits operating above 25 MHz, but by selecting transistors you can still get enough drive to the power amplifier on ten meters. If selecting transistors isn't your cup of tea, simply add an additional driver stage!

When setting up the driver stages, never



C3 Three-section variable, 20 pF per section, (J. W. Miller 1460, third section used in transmitter driver, see fig. 4)

L2 36 μH (Miller 40A335CBI) with 15 turn primary

L3 38 μH (Millar 40A335CBI with 18 turn secondary)

L4 30 turns no. 26 on a ¼" slug-tuned form with 6 turn secondary

Trap 5.2 μH (Miller 40A476CBI)

fig. 5. Receiver front end and high-frequency mixer.

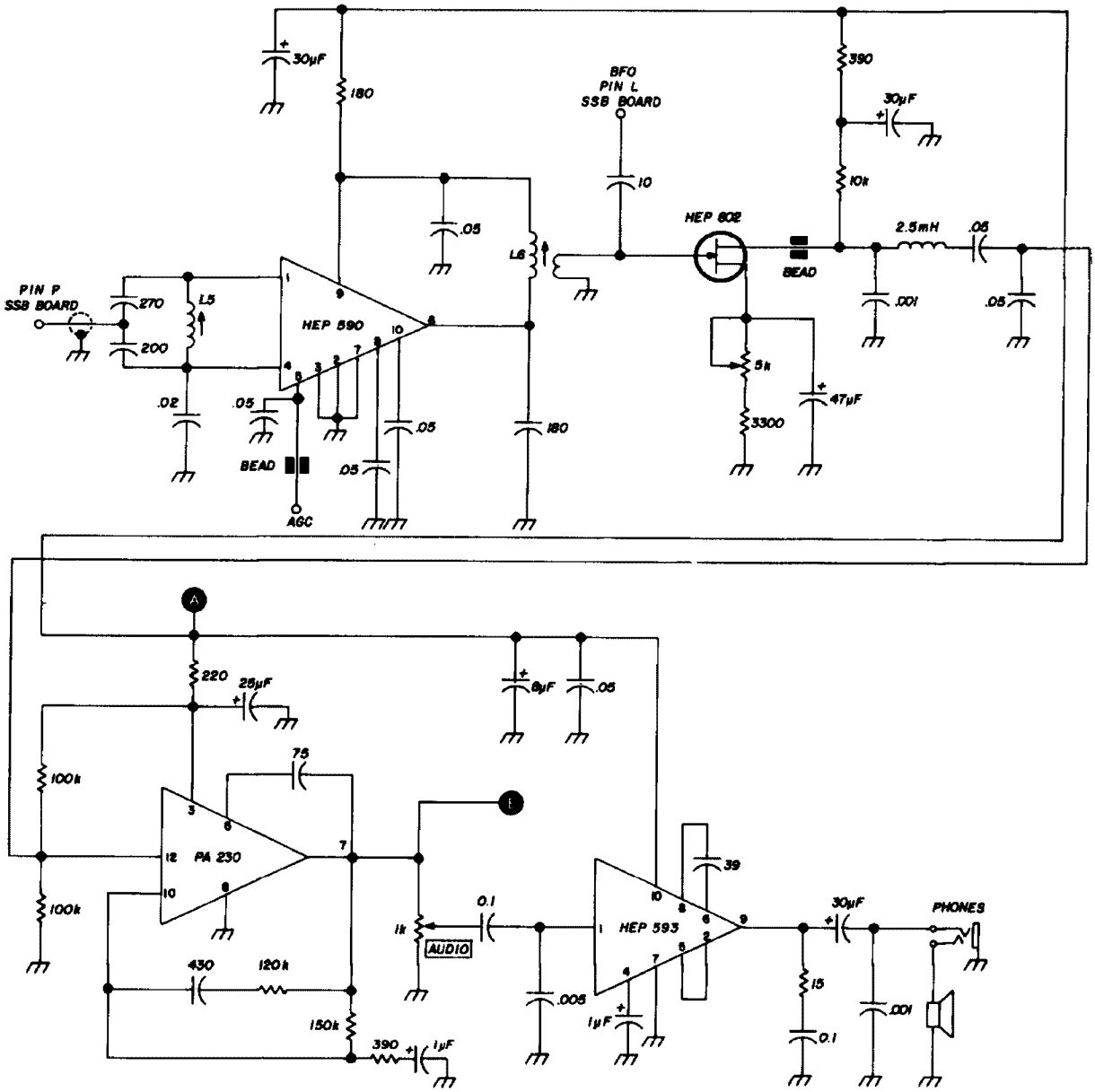


fig. 6. I-f amplifier, product detector and audio power stages. L5 and L6 are 1.6 µH (Miller 40A156CBI), secondary winding on L6 is 4 turns.

turn the power on unless the output tuned circuit is loaded down—either by the next transistor or a resistive load. When the stage is unloaded, the voltage across the high-Q tank circuit may be high enough to burn out the transistor. This is another advantage of the 2N3053: the high maximum-voltage rating allows some mistakes.

When the final driver stage is complete, you should be able to drive it to about 1 watt peak dc input. If you're anxious to put your low-powered creation on the air, you can haywire a T-network together to match the coaxial line to your antenna. With this arrangement on 75 meters one quiet morning, I was able to work Maine, Massachusetts and New Jersey.

receiver section

Although I will discuss the receiver beginning with the mosfet front end, actual construction is simplified if you build the audio stages first and the rf amplifier last! If you build the audio section first, followed by the product detector, i-f amplifier, mixer and rf stage, you can test each circuit as you go along. Then, if you run into a problem, you can fix it before moving on to the next stage.

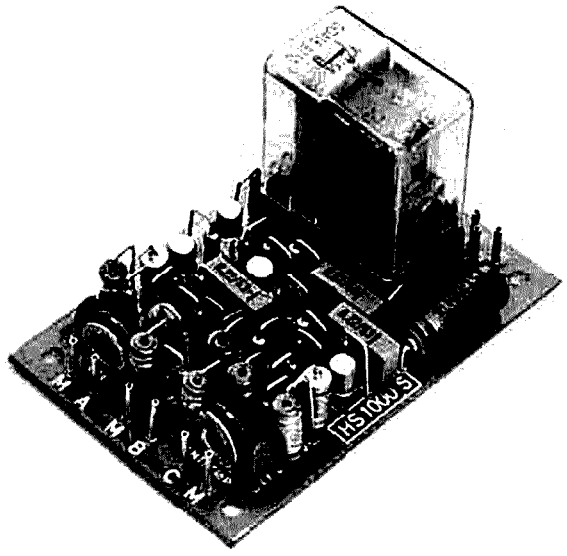
The rf amplifier uses a new dual-gate mosfet, the Motorola MFE3006. This mosfet features good high-frequency (and vhf) performance at reasonable cost. Circuit layout is very uncritical as long as you keep the output isolated from the input. The feed-back capacitance of the transistor itself is very low, so no neutralization *is* required—*if* you have problems with instability, your layout or construction is not up to par.

The high-frequency mixer stage uses a junction fet. The input signal is applied across the gate with local oscillator injection across the source resistor; the output circuit is tuned to 9 MHz. The stage following the mixer is a simple untuned RC-coupled amplifier for driving the 9-MHz crystal filter. Connection to the filter is made through a short length of miniature RG-174/U 50-ohm coax.

i-f amplifier and product detector

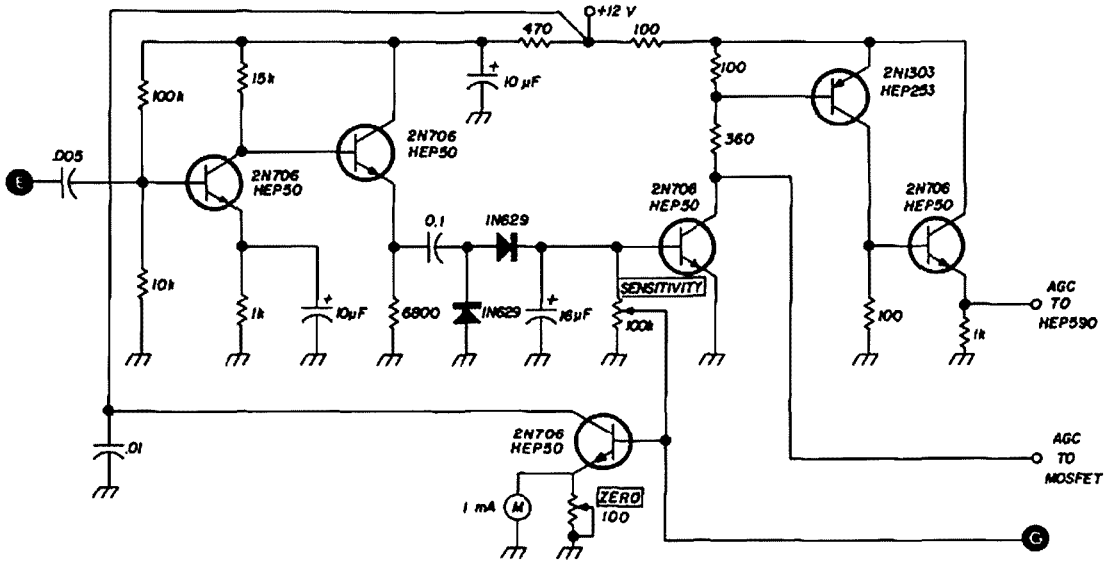
The high-gain 9-MHz i-f amplifier uses an integrated circuit—the HEP 590—a common-emitter, common-base cascode circuit designed for communications equipment. In the circuit shown in **fig. 6,** the input circuit is designed to match the 50-ohm coax line from the crystal filter to the input of the IC.[3] The 9-MHz i-f signal is coupled into the product detector through a 14-turn link.

Although the HEP 590 provides a good deal of gain in a small package, no problems were experienced with instability. Keep in mind when wiring this device into the circuit that pins 2 and 3 should be connected together and grounded; pins 8 and 10 should be bypassed to ground with good quality low-inductance disc bypasses. The gain-bandwidth



Schilling vox/anti-trip module.

fig. 7. Audio-derived agc circuit for the single band transceiver. The s-meter is a Micronta 1-mA instrument available for $2.95 from Radio Shack.



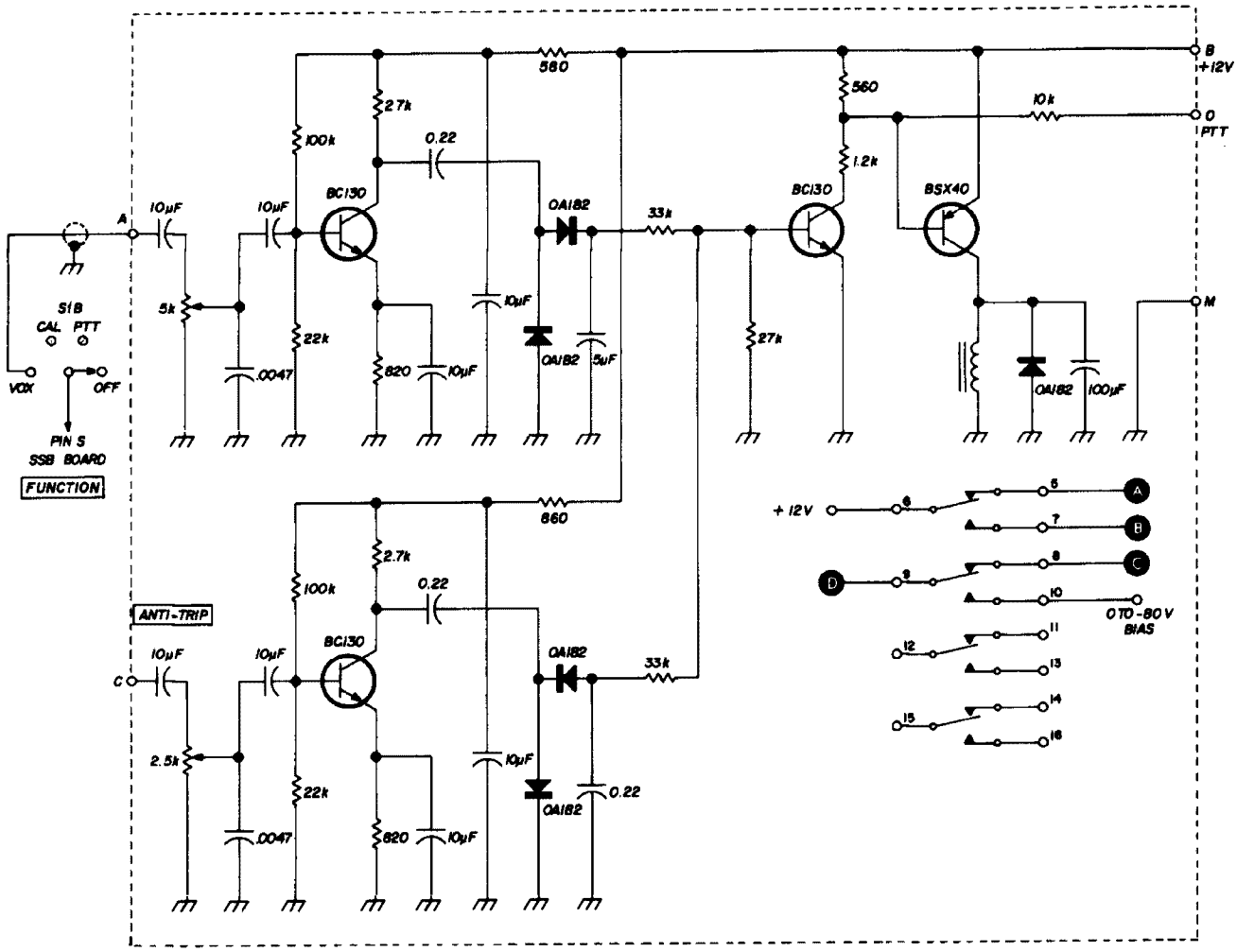


fig. 8. Vox circuitry for the single-band transceiver uses the Schilling HS1000S vox/anti-trip module.

product of the basic IC is in the neighborhood of 2 GHz, so it is imperative that you use good vhf construction practice when wiring it into a circuit.

The product detector uses a junction fet with both signal and bfo injection across the gate. The optimum bias point for the transistor is set by the pot in the source lead. With the bfo turned off, you should get practically no signal through to the audio stages. The bfo injection capacitor is adjusted for maximum gain with minimum noise contribution. If too much bfo injection is used, the product detector will be very noisy. Some fet's work better than others in this circuit; the zero-bias current range for the HEP 802 is 2 to 20 mA—devices with the lower values work best. The product detector in the transceiver shown in the photographs consumes 500 µA with no bfo injection and about 600 µA with the bfo turned on.

The receiver audio section uses two integrated circuits—a PA230 preamplifier and HEP 593 audio power stage. Output power is greater than 1 watt with a 12-volt supply. The bypass and coupling capacitors in the audio section were chosen to limit the frequency response to 300 to 3000 Hz, the range required for communications work.

When wiring the two audio IC's be very careful to use short direct leads. The individual transistors in the HEP 593 audio power IC have excellent vhf response, so circuit layout is very important. Instability in this stage can be caused by excessive lead inductance or stray capacitance; parasitic oscillations may even cause rf heating and eventual device burnout. The circuit shown in **fig. 6** includes

all the necessary bypasses.

The RC bypassing network from pin 9 to ground should be placed right next to pin 9. This circuit eliminates vhf instability caused by the inductance of the speaker leads. The power supply line should be bypassed by a good quality vhf bypass capacitor directly from pin 10 to ground. In addition, leads to pins 7, 9 and 10 should be as short as possible and the input should be isolated from the output through good layout and short leads.

For maximum audio output with a 12-volt supply, the HEP 593 should be used with an 8- or 16-ohm speaker. Although the circuit will deliver usable power into a 3.2-ohm speaker, it has to work a lot harder to do it, and harmonic distortion is much higher. For best results, use a small transistor auto-radio replacement speaker; these usually have a higher voice-coil impedance.

power supply

One of the design goals for this transceiver was a built-in power supply for both mobile and fixed station use. This is provided by the circuit shown in **fig. 10.** The transformer is a standard commercial off-the-shelf item manufactured by Thordarson. The high-voltage secondary is used with a voltage-doubling circuit that provides +600 volts for the plate and +300 volts for the screen of the 6883 power amplifier. A negative bias supply is provided by a shunt rectifier circuit off the secondary winding.

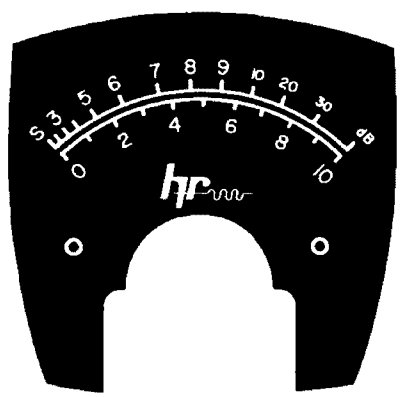
When the transceiver is operating from the 117 Vac line, the regulated power supply for the solid-state circuitry is developed from a full-wave bridge across the 12-volt winding on the transformer. When the power supply is connected to the 117-volt line, the dc-to-dc converter transistors must be disconnected from the power transformer. This is accomplished automatically by the two different power plugs shown in **fig. 10.**

Total current drain is very low in all modes. The receiver requires from 60 to 100 mA depending upon the audio level. Current drain from a 12 volt battery while transmitting depends on loading. However, when the final is loaded up to 100 mA, total input current from the 12-volt supply is 6 amperes.

construction

All of the transceiver circuits except the power amplifier stage and power supply are built on pieces of copperclad board. Although printed circuits were not used in the transceiver shown in the photographs, they would be ideal for this purpose and would result in a somewhat more rugged unit. I doubt that reliability would be improved much since the unit shown here has been extremely dependable.

The transceiver circuitry is built on four different boards: transmitter mixer and driver stages; receiver front end, mixer and crystal-filter driver; receiver i-f, audio and agc; and vfo. The 9-MHz sideband generator and vox boards are mounted on top of the 4x7x1½-inch chassis on the right-hand side of the unit.



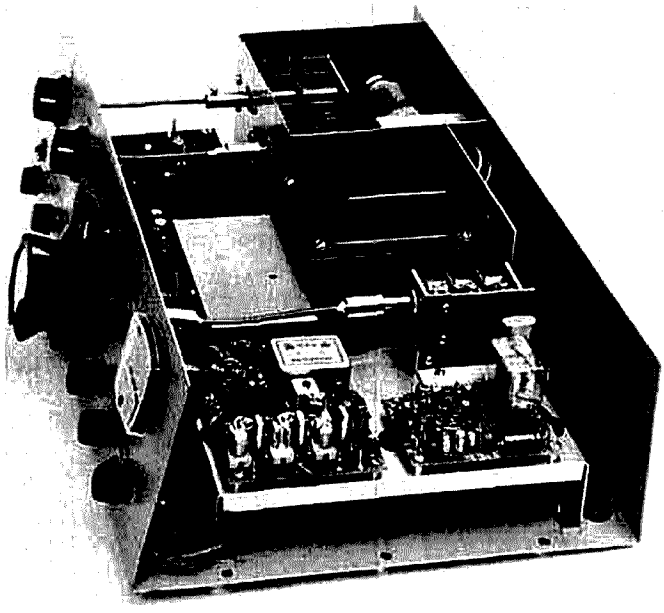
New dial face for the s-meter.

The receiver front end, mixer and crystal filter driver board is underneath as are the transmitter and mixer stages.

The vfo board is installed inside the 2x2x4-inch minibox. One side of this box is bent back at a 60° angle to conform with the front panel of the "Tilt-a-View" cabinet (Bud TV-2155). The sloping front panel requires that the vfo tuning capacitor be mounted on an angle so the shaft is perpendicular to the front panel; the shafts from the final-amplifier tuning capacitors use flexible couplers to provide the necessary displacement. The other controls are mounted on the front panel. The end result is a miniature, modern-looking package that can be carried around in a brief case.

Although the sloping front panel wastes some space, there is still plenty of room to spare. Although it's difficult to see in the photographs, the area under the chassis is almost completely unused. There's plenty of room for more circuitry—such as an additional vfo, a noise blanker, sidetone oscillator, speech processor or even an electronic keyer!

If you want to use a more conventional cabinet, there are any variety available. However, all of the modern-looking ones are so large you could package this transceiver and have room left over for a high-powered linear amplifier. This might not be such a bad idea for the home station, but for mobile and portable operation, the smaller size and weight of the package shown in the photos is more desirable.



Internal layout of the transceiver. Tho 9-MHz ssb generator, vox board, receiver rf amplifier and transmitter drivers are mounted above and below the chassis in the foreground. Vfo box is in front of the power transformer. Heat sink for tho power transistors is behind the rear panel. I-f amplifier, audio power and agc stages are built on the board in front of the power amplifier compartment. Flexible shafts are lengths of speedometer cable epoxied to short pieces of ¼" tubing.

The only other construction point of importance concerns the vfo tuning knob. The arrangement shown in the photo consists of two back-to-back vernier drives*—one on each side of the front panel. Each drive has a 6:1 reduction ratio. When they are connected in tandem as they are here you have a choice of either 6:1 or 36:1 reduction. The 6:1 (with the larger knob) is ideal for moving around ihe band; the 36:1 is perfect for tuning sideband.

The tuning dial is mounted on the reduction drive that is mounted inside the cabinet. The escutcheon was cut out from a piece of quarter-inch aluminum, filed to shape with a hand file and painted black.

alignment

Before lining up the transceiver circuits, it's a good idea to make sure the 9-MHz excited module is properly aligned. This is a simple process with the directions furnished by Spectrum International and consists primarily of setting the carrier oscillators in proper relationship to the filter passband.

Vfo alignment is essentially a process of setting up the trimmers in the tuned circuit so the 50-pF variable will tune the oscillator over ihe desired range. For 80 and 10 meters, a full 500-kHz range is required to cover the band; on the other h-f bands, slightly less coverage will give you more bandspread—practical ranges are 300 kHz for 40, 350 kHz for 20 and 450 kHz for 15. If you are not interested in CW work, you may want to limit vfo coverage to the phone section of the band. For 10 meters you might even want to increase the vfo range to a full megahertz, but ssb tuning will be a little rough with this much coverage.

Regardless of the range you want, the vfo alignment procedure is basically the same. With the constants shown in **fig. 3** the basic tuning range is from about 4095 to 5505 kHz. With the 50-pF variable fully meshed, set the oscillator frequency to approximately 5500 kHz with the 20-pF trimmer; when the variable is fully open, the frequency should be approximately 5000 kHz. If you can't get

* Jackson Brothers type 4511 DAF. $1.50 each from Arrow Electronics Inc., 97 Chambers Street, New York, New York 10007.

the full 500 kHz range, replace the 75-pF shunting capacitor with the next smallest value. If the tuning range is larger than you want, increase the value of the shunt capacitor.

If you change the range within a few kHz, the inductance should require no adjustment. However, if you want to make a big reduction in vfo range, you may find it necessary to remove or add several turns to the toroid inductor. If you leave 4- or 5-inch pigtails on the inductor when winding it, you'll have plenty to work with when one stage at a time, peaking each tuned circuit as you go. Be sure to decrease signal generator output so you don't overload the receiver stages. Finally, hook the 9 MHz source to the antenna jack and adjust the 9-MHz trap for minimum signal feedthrough.

To line up the rf amplifier, tune the signal generator to the low edge of the amateur band, fully mesh the variable capacitor and tune coils L2 and L3 for maximum. Move the signal generator up to the top edge of the band, unmesh the variable and peak up the trimmers on the variable. After

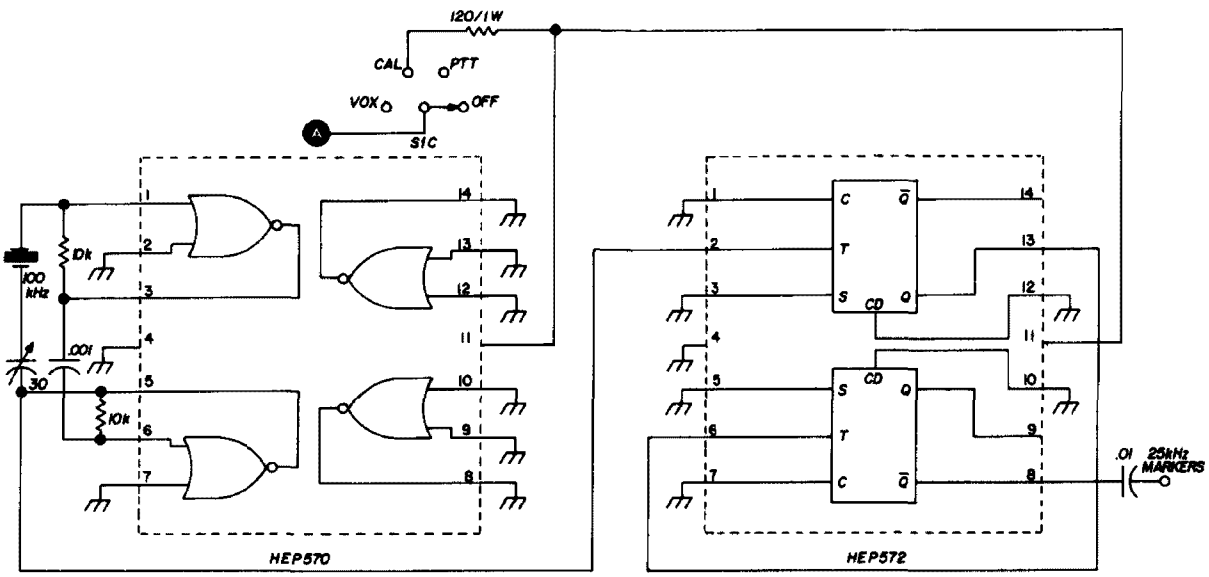


fig. 9. Crystal calibrator is mounted on the same board as the receiving rf amplifier.

setting up the frequency range. A little juggling of values may be required to cover the precise range you need, but it is not difficult nor time consuming. Remember that the inductor essentially determines the center of the frequency range; the shunting capacitor determines how much the frequency is changed by the air variable.

Receiver alignment is next and consists of peaking up the i-f amplifier and receiver mixer tuned circuits. If you built the unit from the audio section back, this is probably already done. If it isn't, inject a 9 MHz signal to the gate of the product detector through a 1000-pF capacitor. Tune the signal generator around until you get a beat note. Now move the 9 MHz source toward the rf amplifier, several tries it will be found that neither the lower-frequency inductor settings nor the upper-frequency capacitor settings will further improve the signal.

Before lining up the transmitter stages, remove the final power amplifier tube from its socket. A sweep generator is a big help in lining up the broadband transmitter drivers, but it isn't absolutely necessary. If you have a sweep generator, set it to sweep from 5.0 to 5.5 MHz; adjust the output to approximately 1 volt p-p and connect it to the transmitter balanced mixer (vfo injection mixer if you use one). Turn the exciter module to the CW position and insert sufficient carrier to provide a respectable trace on the scope; now align each of the

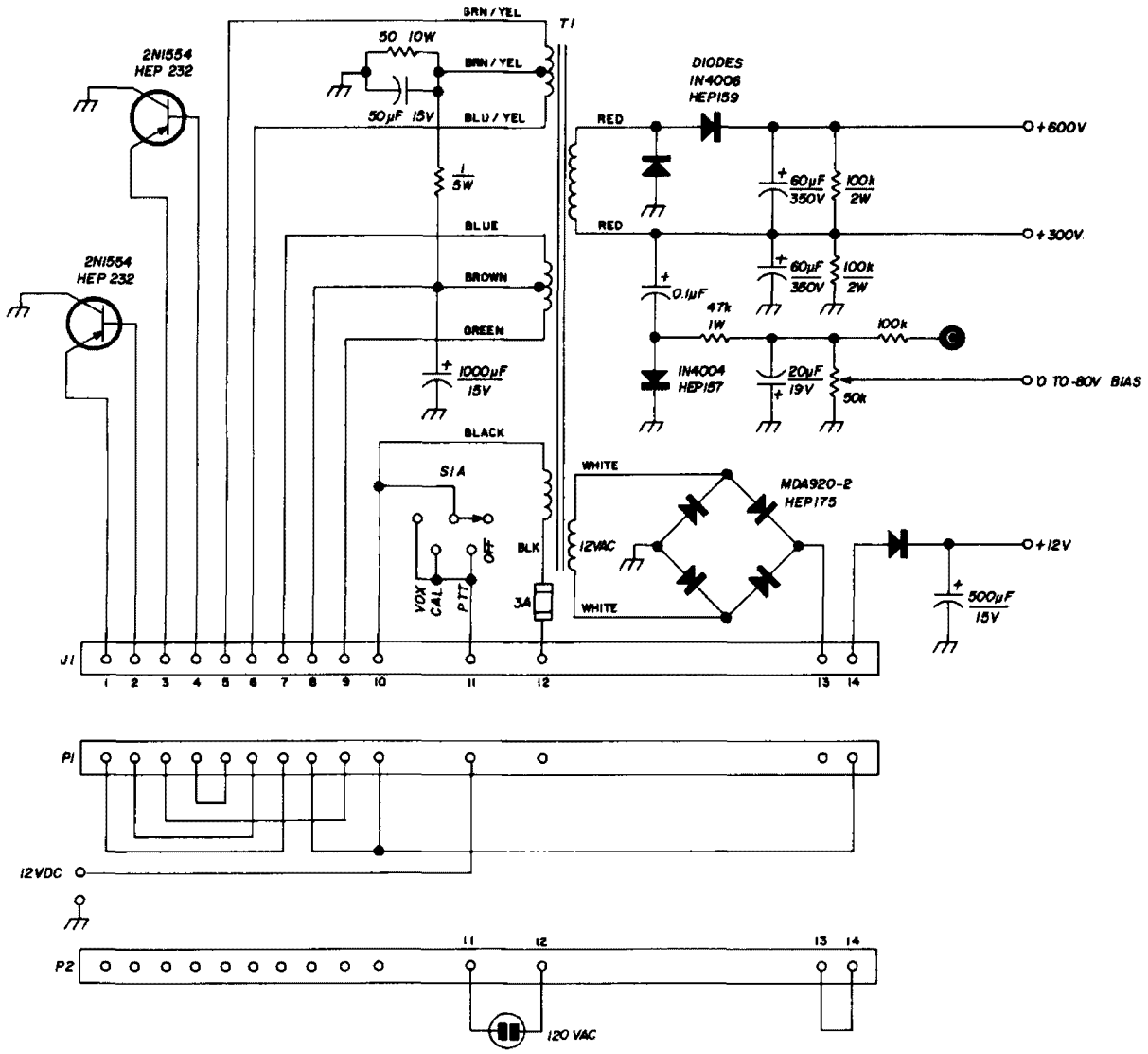


fig. 10. Power supply for the single-band transceiver operates either 120 Vac or 12 Vdc. Transformer T1 is a Thordarson TR 294. The 2N1554 power transistors must be mounted on heat sinks.

broadband transmitter stages for flat response over the amateur band.

If you can't beg or borrow a sweep generator, set the vfo about 10 percent in from the lower edge of the band and peak each of the stage **input** circuits; then tune the vfo in about 10 percent from the top edge of the band and peak each of the **output** circuits. After going through all the stages several times, there should be no further improvement in signal at either end of the band. As a final check, the output signal (measured at the grid pin of the power amplifier tube) should vary less than 1 dB when the vfo is tuned through its range.

When all the solid-state circuits are aligned, plug in the power amplifier. After it has warmed up, adjust the bias for 25 mA no-signal plate current. With the exciter control in the tune position, adjust carrier insertion and drive, then dip the final. If everything is in order, you should get about 50 watts input to the dummy load.

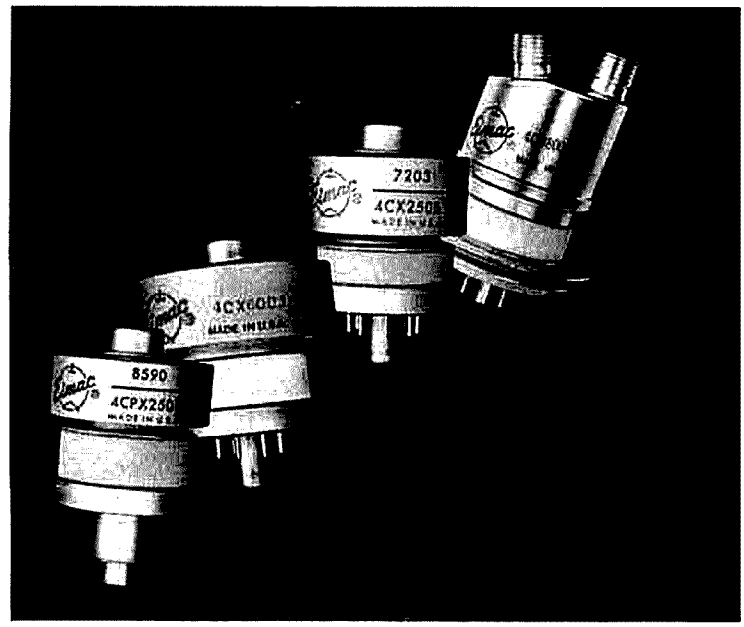
references

1. J. R. Fisk, W1DTY, "Stable Transistor VFO's," *ham radio*, June, 1968, p. 14.

2. R. C. Hejhall, "Getting Transistors into Single-Sideband Amplifiers," Motorola Application Note AN-150, 1965.

3. J. Robertson and B. Welling, "An Integrated-Circuit RF/I-F Amplifier," Motorola Application Note AN-247, 1968.

ham radio



external-anode tetrodes

This is a handy reference source, complete with quantitative data, on these popular rf generators

William I. Orr, W6SAI, Manager, Amateur Service Department, Eimac Division of Varian, San Carlos, California 94070

The 4X150A external-anode tetrode was developed by Eimac back in 1947. This highly efficient, vhf-rated transmitting tube is the forerunner of a whole family of similar tubes that have evolved over the years since the war. The tube is now manufactured by a host of companies, but the original design is still going strong, and it's still the choice for a reliable rf power source with many vhf enthusiasts.

During the past few years a bewildering number of "4X-..." tubes have appeared. Each has characteristics slightly different from the original 4X150A. I'd like to discuss this family of tubes and identify, once and for all, the basic characteristics of each type. Amateurs interested in these tubes will then have a ready reference for future applications.

the original 4X150A

Let's take a look at how the nomenclature describes the tube. The numeral **4** signifies four active elements: cathode, control grid, screen grid, and anode—a tetrode. The **X** indicates an external anode (forced-air cooled), **150** indicates the plate

Eimac	EIA	Po Watts	Ep Volts	Fmax MHz	Heater V/A	Base	Figure of Merit	Notes
					4X150A Family			
4X150A (old)	—	150	1250	500	6.0/2.6	9-pin	86	old style anode; weight: 5.2 ounces
4X150A	7034	250	2000	500	6.0/2.6	9-pin	86	new style anode; weight: 4 ounces
4X150D	7035	250	2000	500	26.5/0.55	9-pin	86	aircraft version of 4X150A
4X150G	8172	150	1250	1200	2.5/6.25	coaxial	56	uhf and video service
4X150R	8296	250	2000	500	6.0/2.7	9-pin	73	ruggedized 4X150A/7034
4X150S	8297	250	2000	500	26.5/0.56	9-pin	73	ruggedized 4X150D/7035
					4X250B Family			
4X250B	—	250	2000	500	6.0/2.6	9-pin	85	ceramic shell, glass-based 4X150A
4X250F	—	250	2000	500	26.5/0.56	9-pin	85	aircraft version of 4X250B
4CX250F	7204	250	2000	500	26.5/0.56	9-pin	85	all ceramic 4X250F
4CX250B	7203	250	2000	500	6.0/2.6	9-pin	85	all ceramic 4X250B
4CX250K	8245	250	2000	1200	6.0/2.6	coaxial	54	uhf and video service
4CX250M	8246	250	2000	1200	26.5/0.56	coaxial	54	aircraft version of 4CX250K
4CX250R	7580W	250	2000	500	6.0/2.6	9-pin	81	ruggedized 7580
7580W	7580	250	2000	500	6.0/2.6	9-pin	82	high-perveance 4CX250B
4W300B	8249	300	2000	500	6.0/2.6	9-pin	85	water-cooled 4X250B

table 1. External-anode tetrode characteristics.

dissipation in watts, and **A** signifies the first production version.

Recent productions of the 4X150A have included a radically new brazed anode structure having a complex arrangement of cooling fins. The new design allows increased plate dissipation ratings. The designation of this version is 4X150A/7034. It is operationally equivalent to the 4CX250B at frequencies below 150 MHz.

how to identify them

The different versions of the 4X150A are not easy to tell apart. Some transitional 150-watt tubes have the improved cooling fins but not the brazed anode structure. In general, the old- and new-style 4X150A's may be distinguished by weight. The 150- and 250-watt tubes weigh approximately 5 and 4 ounces respectively; their respective anode diameters are 15/16 and 13/16 inch. Interestingly enough, plate dissipation has been increased with reduced weight and size.

High-voltage heater versions of the 4X150A were introduced as the 4X150D and 4X150S (ruggedized). These new tubes, plus the redesigned descendants of the 4X150A, bear alternate Electronic Industries Association (EIA) nomenclature. This consists of four-digit numbers in the seven- and eight-thousand series. Thus the EIA 7034 is the 4X150A, etc.

The 4X150G (2.5-volt heater) and its offspring, 4CX250K (6-volt heater), have coaxial terminations. They are designed for internal cavity operation at frequencies into the gigahertz region.

Eimac	EIA	Po Watts	Ep Volts	Fmax MHz	Heater V/A	Base	Figure of Merit	Notes
					4X150A Family			
4CX300A	8167	300	2500	500	6.0/2.7	breech-block	52	ceramic-metal ruggedized
Y-180	—	300	2500	500	6.0/2.7	breech-block	52	nickel-rhodium plated 4CX300A
4CX300Y	—	—	2500	500	6.0/3.4	breech-block	53	high-plate current version of 4CX300A
4CX350A	8321	350	2000	500	6.0/3.0	9-pin	143	high transconductance, high current 4CX250B
4CX350F	8322	350	2000	500	26.5/0.57	9-pin	143	aircraft version of 4CX350A
4CX125F	—	125	2000	500	26.5/0.57	9-pin	52	aircraft version of 4CX125C
					special versions			
4CN15A	—	15	2500	500	6.0/3.0	breech-block	52	low-duty pulse work or heat-sink cooling
4CN15L	—	15	2000	—	2.1/7.5	9-pin	—	quick-heat cathode
4CS100L	—	100	2000	—	2.1/7.5	9-pin	85	quick-heat cathode, heat-sink cooling
4CX125C	—	125	2000	500	6.0/2.7	breech-block	52	horizontally finned 4CX300A
4CX125F	—	125	2000	500	26.5/0.56	breech-block	52	identical to 4CX125C except for filament voltage
4CX250L	—	250	2000	—	2.1/7.5	9-pin	—	quick-heat cathode
4CPX250K	—	250	5500	500	6.0/2.7	coaxial	85	pulse rated 4CX250K

the 4X250B family

The 4X250B features ceramic insulation. It evolved from the glass-insulated 4X150A. Early versions were made with a ceramic outer cylinder and a glass base; later designs are all ceramic and are called 4CX250B. These 4X-series tubes are rated at 250 watts with a maximum plate voltage of 2000.

The latest offspring of the 4X150A is the popular, rugged 4CX300A. This Ceramic and metal tetrode operates at plate voltages up to 2500. Special-purpose versions are currently in production. Among these is the 4CX300Y, which should be of interest to ssb amateurs and others contemplating new equipment. The 4CX300Y resembles the 4CX300A in appearance, but has a heavy-duty heater (6.0 volts at 3.4 amperes). This permits unusually high values of plate current.

external anode tetrodes in grounded grid?

Modern, high-gain external-anode tetrodes don't perform well in the conventional class-B grounded-grid arrangement for several reasons. The external-anode tube is characterized by high perveance, together with extremely small spacing between grid bars and between grid and cathode. Thus while performing excellently as grid-driven tetrodes, they are not suitable for grounded-grid operation as such.

For proper operation, the screen requires much higher voltage than the control grid. Older tetrodes having lower gain tend to have more equal balance between absolute grid and screen currents. When these electrodes of the newer, high perveance, external-anode tetrodes are tied together, the control grid tends to draw tre-

mendous currents, and there is grave risk of destroying the grid. Peak grid current, for example, in a 4X150A operated in grounded grid can easily be **twice** the value of peak plate current.

It's permissible, however, to operate the external-anode tube as a cathode-driven tetrode, with the grid and screen at rf ground, but operating at the normal dc potentials. Grid dissipation under these conditions is minimal, and stage gain is greatly increased. Screen dissipation is nearly the same as in the tetrode connection. Greater stage gain can be obtained with this circuit, because the driver doesn't have to supply large screen and grid losses. Excess drive power should be absorbed in a resistive load.

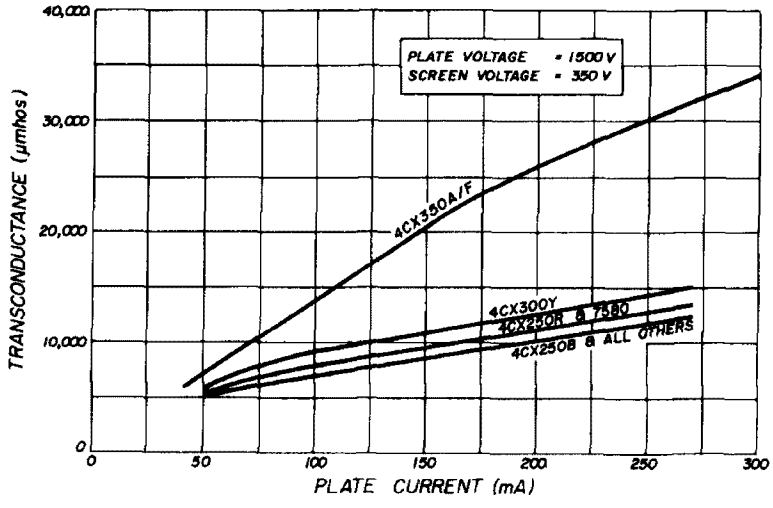


fig. 1. Mutual transconductance vs plate current for several external-anode tetrodes. The 4CX350A and 4CX350F have about twice the transconductance as the rest of the line; the 4CX300Y has approximately 30% higher transconductance than the "standard" tubes while the 7580 and 4CX250R/7580W are approximately 20% higher.

the tube socket

In addition to permitting connections to the elements, the socket for external-anode tubes conducts heat away from the tube stem and, in some cases, serves as a capacitive screen bypass.

Complete Air-System socket assemblies for all noncoaxial-based external-anode tubes are available. These consist of socket and air chimney (**table 1**). The sockets are designed to permit air to be directed axially on the tube base, past the base to the envelope, and then over the anode cooler. Use of other than the Air-System socket with external-anode tubes is not recommended, because tube temperature can't be adequately controlled. A receiving-type loctal socket with 4X150A-style tubes is emphatically **not** recommended. Dangerously high stem temperatures will occur from the filament heat unless the base structure is cooled by an air blast. The solid construction of the simple receiving-type loctal socket blocks the flow of air around the tube stem.

I'd like to emphasize that heater voltage on the 6-volt external-anode tetrodes is 6.0 volts ±5 percent, **not** 6.3 volts. What this means is that the tube will operate at 6.3 volts, which is the upper limit of the specification (+5 percent), but for longest heater life, the voltage **should not** exceed 6.0 volts rms.

Anyone using these tubes, or any transmitting tube for that matter, should measure heater voltage **at the tube heater terminals.** The meter should be calibrated against a one-percent laboratory instrument, and the measurement should be made under full-load, key-down conditions.

You would do well to take a tip from broadcast and TV stations. Some of the large tubes in these installations cost around $1,000.00 to replace, and no maintenance engineer is about to face the embarrassment of having a final amplifier tube go west because of high heater voltage. At KFMB-TV, for example, the senior transmitter engineer (W6QCN) says that their

big tubes have a meter in the input to the heater voltage source, which is regulated to supply the exact voltage at the tube heater terminals.

the figure of merit

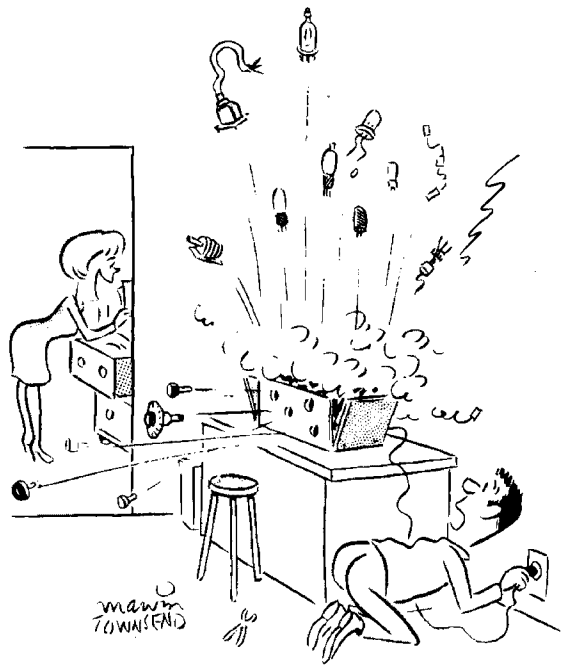
A graphical presentation of the mutual conductance for external-anode tubes as a function of plate current is given in **fig. 1.** The 4CX350A and 4CX350F have about twice the g_m of other tubes in this class. The 4CX300Y has about 30 percent higher g_m, and the 7580 and 7580W/4CX250R are about 20 percent higher than the 4X150A. Mutual conductance (or transconductance, as it's sometimes called) is a figure of merit of a particular tube. Also known as the gain-bandwidth factor, this quantity is

$$\text{Figure of merit} = \frac{g_m}{2\pi C_t}$$

where C_t is the total input and output capacitance of the tube.

The figure of merit is a relative number and should not be interpreted as an absolute value. C_t is an average, taken from a rumber of typical tubes. Highest figure-of-merit values are reached by a combination of high g_m and low interelectrode capacitances. The 4CX350A and 4CX350F, which have the highest g_m and reasonably low interelectrode capacity, have the highest figure of merit. The coaxial- and breechlock-based tubes, with their higher capacitances, appear to have lower figure-of-merit values. However, the coaxial-based tubes perform more efficiently at the higher frequencies, and they are especially designed for cavity operation.

ham radio



**"Good news, dear!
I've finally found the instructions
for that kit you're trying to assemble."**

lighthouse tubes for uhf

Planar triodes or lighthouse tubes are well suited for amateur use in the uhf spectrum; various surplus versions, such as the 2C40, 446B and 2C39 have been used at frequencies up to 2400 MHz. A relatively new member of the family, the 3CX100A5, is available for improved service at uhf.

The 3CX100A5 is relatively unknown to the amateur fraternity, but its older relative, the 2C39A, has been long a favorite uhf tube of the "surplus hounds." The 2C39A and its twin, the 7289, fit into the amateur picture nicely as amplifiers, doublers or triplers in the range from 1000 to 3000 MHz. They are not expensive, and various glass versions of the older 2C39/2C39A/2C39WA can often be obtained at give-away prices on the surplus market. The 3CX100A5 is an improved, modern version of this old World War II tube, dressed up in a brand-new ceramic and metal envelope.

Performance of the 3CX100A5 as a grounded-grid uhf power amplifier and multiplier is shown in **fig. 1, 2** and **3.** Because of the high power gain of the tube,

grounded-grid circuitry is desirable since intercoupling between the input and output circuits is reduced to a minimum and neutralization is not required. The graphs may be used to estimate the performance of the 2C39 glass family of tubes by noting that the useful power output of this style tube will be somewhat less (depending upon the frequency)—up to 25 percent at 2.5 GHz.

A variety of 3CX100A5 tubes were run in a coaxial cavity capable of tuning from 1000 to 3000 MHz, and a series of measurements was made with a representative sample of production tubes operating as amplifiers, doublers and triplers. Drive and output power were carefully measured for each test, and appropriate filters were used to eliminate feedthrough and harmonic output. When a tube is operated as a doubler or tripler, feedthrough power is at the driving frequency and is undesired. In practice it is necessary to eliminate this power from the output circuit of any grounded-grid frequency multiplier; high-Q tuned circuits or wave filters will do the job.

amplifier performance

A graph of average tube performance as a grounded-grid amplifier is shown in **fig. 1.** Grid drive, grid bias and plate loading were adjusted to provide maximum power output while maintaining plate current at 100 mA. Plate voltage for these tests was 1000 volts, and grid bias was −30 volts.

At 1300 MHz, the 3CX100A5 is capable of about 47 watts power output at an efficiency of 47%. The power gain of the tube is 8 decibels, indicating a required drive level of about 7.5 watts as measured at the input to the cathode cavity.

Power output gradually decreases as the frequency of operation is raised. At 2400 MHz power output (at 100 watts input) drops to 25 watts, and grid driving power increases to 10 watts. Power gain at this frequency is 2.5. A power output and efficiency curve for the 2C39A over the same frequency range is shown by the dotted line.

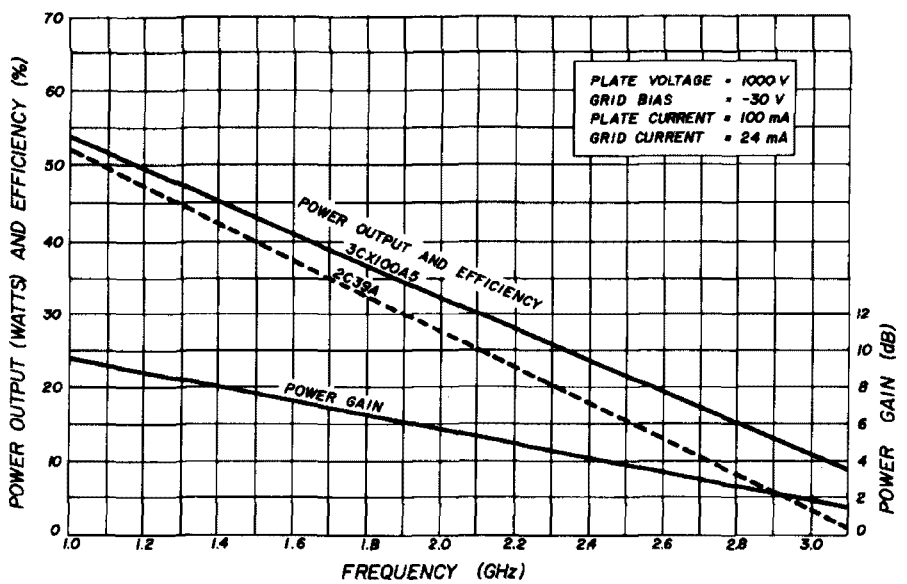


fig. 1. Typical power output, gain and efficiency of 3CX100A5 and 2C39A grounded-grid amplifiers.

doublers and triplers

After completing the amplifier tests, the same tubes were operated as frequency doublers in gain and power output tests. Excitation was applied and the operating conditions and tank circuits adjusted for maximum power output while maintaining 100 mA plate current. Plate supply voltage was 1000 volts, so the operating con-

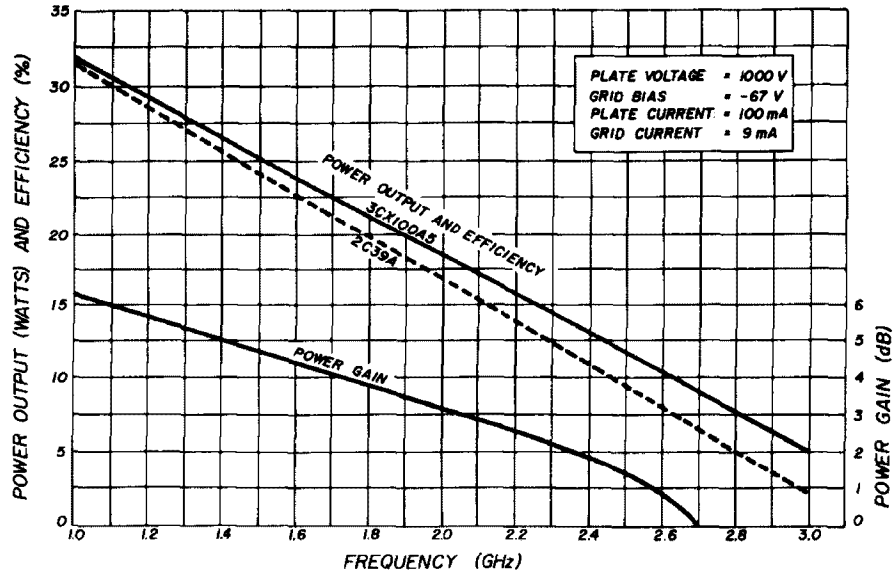


fig. 2. Typical power output, gain and efficiency of grounded-grid doublers.

ditions represent 100 watts input.

At 1300 MHz, with 650-MHz drive, power output was 27 watts; this is a circuit efficiency of 27 percent. Drive power of 8 watts was required, a power gain of 5.3 dB. At 2400 MHz, power output was 13 watts with 9 watts drive at 1200 MHz. Power gain as a doubler drops to unity at about 2700 MHz although the 3CX100A5 is still useful as a doubler at this frequency since 10 watts output can be obtained. This data is plotted by the solid lines in **fig. 2;** 2C39A performance as a doubler is plotted by the dotted line. The same tubes and general test techniques were used to determine the operating parameters of the 3CX100A5 as a frequency tripler. Drive power was applied at one-third the output frequency and the circuit was adjusted for maximum power output. At 1300 MHz, 17 watts of power were obtained with about 10 watts drive at 435 MHz. At 2400 MHz, power output was about 8 watts with 12 watts drive at 800 MHz. Power gain as a tripler drops to unity just above 2700 MHz but useful power output is still available, although efficiency is very poor.

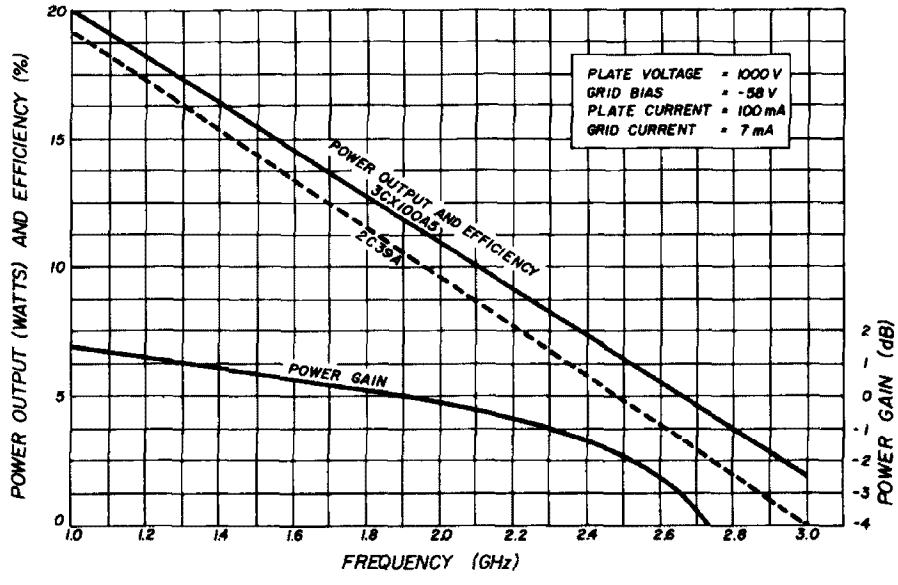


fig. 3. Typical power output, gain and efficiency of grounded-grid triplers.

table 1. Planar triode tubes.

2C39A	Glass or ceramic construction	Eimac, Mschlett
2C39WA	Glass or ceramic construction	Eimac, Machlett
2C41	Not interchangeable physically with 3CX100A5 or 7289	Machlett
3CX100A5	All ceramic construction	Eimac, Machlett, GE
3CX100F5	Identical to 3CX100A5 except heater voltage is 26.5 volts	Eimac
381	3CX100A5 type, rated for pulse service	Eimac, Machlett
6897	Not interchangeable electrically with 3CX100A5 or 7289 in most uhf sockets	GE
7289	Identical to 3CX100A5	Eimac, Machlett, GE

A socket for the 3CX100A5 is a rare bird since it is an integral part of the resonant cavity in most equipment. However, collet rings for the tube are available from the Instrument Specialties Company or Braun Tool and Instrument Company. A complete socket assembly can be purchased from Jettron Products.*

conclusion

The planar triode tubes of the 3CX100A5 family perform well in the frequency range encompassing the 1215-MHz and 2400-MHz amateur bands. Efficiency is good, considering the frequency of operation. F. E. Terman indicates[1] that a doubler will provide about 65 percent of the power output of a straight-through amplifier, and a frequency tripler will provide about 40 percent of the power output of the amplifier. These figures agree closely with the data shown here. Two recent articles[2,3] show that these tubes can put quite a dent in the 1215-MHz band.

references

1. F. E. Terman, "Radio Engineering," McGraw-Hill, New York, 1962.
2. P. Laakmann, WB6IOM, "Cavity Amplifier for 1296 Mc" *QST*, January, 1968, p. 17.
3. P. Laakmann, WB6IOM, "Kilowatt Amplifier for 1296 MHz," *ham radio*, August, 1968, p. 8.

* Instrument Specialties Company, Little Falls, New Jersey; plate collet, 97-70; grid collet, 97-72; cathode collet, 97-76; and filament collet, 97-80.

Braun Tool and Instrument Company, 140 5th Avenue, Hawthorne, New Jersey; plate collet, 134-53; grid collet, 134-51; cathode collet, 165.

Jettron Products, Inc., 56 Route 10, Hanover, New Jersey, can supply complete sockets.

Bob Sutherland, W6UOV

water cooling the 2C39

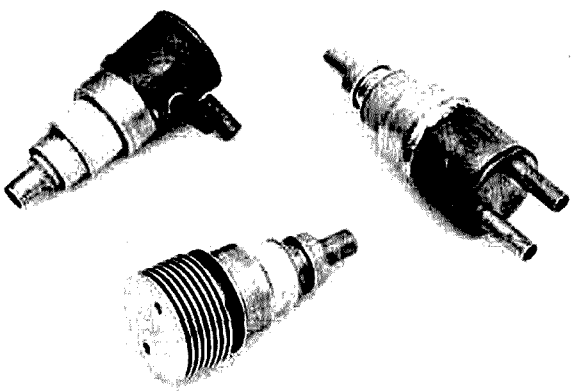
There is a gradual movement of amateur radio to higher and higher frequencies where there is more spectrum and room for interesting experimentation. Probably the most popular tube for amateur operation on 1296 and 2300 MHz is the erstwhile 2C39 or a modern member of the same family. Originally air cooled and rated at 100 watts plates dissipation, by using water cooling these tubes can safely dissipate 150 to 200 watts without noticeably shortening tube life. If you want to build efficient high-power and **quiet** amplifiers and multipliers for these bands with 2C39-type tubes, water cooling is the answer.

water jacket

First of all, slightly loosen the two Allen screws that hold the finned heat sink in place and pull it away from the tube body. Even with the screws loosened, pull-

ing the heat sink off may be difficult because the heat sink is aluminum and the tube anode is silver-plated copper. I've found that cooling the whole works in a freezer sometimes helps; sometimes lubricating with *acetone* will help. Not all 2C39 manufacturers use Allen screws to hold the heat sink in place; the heat sinks screw onto the anode of some tubes, but this type can be removed as well.

Watercooled 2C39's. Side-inlet versions are used in a UPX-4; top-inlet types are used in a WB6IOM two-tube 1296-MHz amplifier.



The water jacket is made from a 1-inch length of 1-inch diameter (ID) copper or brass tubing. Punch out a 1-1/8-inch disc from 1/16-inch copper or brass sheet with a punch. Next, drill two 1/4-inch holes for the water inlet and outlet tubes 7/16 inch apart. These can be located in one side of the water jacket or on top, depending upon the application. Cut two 1/4-inch OD copper or brass tubes about 3/4-inch long; these tubes should have an inside diameter of 3/16 or more.

Fit the tubes into the 1/4-inch holes in the jacket so that their ends are flush with the inside surface of the jacket; silver solder the four pieces together—jacket, disc and two tubes. After cleaning the completed water jacket to a shiny finish, soft solder it to the 2C39 anode with a small torch. Be sure to keep the flame entirely on the water jacket to prevent damage to the 2C39.

coolant system

Cooling water is run into the 2C39's with 3/16-inch inside-diameter vinyl or tygon tubing. If you use more than two tubes in the system, it's a good idea to use intake and exhaust manifolds for splitting and recombining the water. The manifolds are made from 8-inch pieces of 1-1/8-inch OD copper or brass tubing.

Quarter-inch holes—spaced 1 inch apart—are drilled along the sides of the tubing; small brass tubes, 1/4-inch OD and about 3/4-inch long are pushed into these holes. One end of the manifold tube is capped with a 1/16-inch thick brass disc and the other end is fitted with standard garden-hose fittings. Then all the pieces are silver soldered together.

Half-inch hose is used for the main lines to and from the pump. The pump is a "Little Giant" submersible pump that is designed for small backyard waterfalls; this inexpensive little pump doesn't make any noise while it's pumping. Be sure to ground the manifolds because the cooling water has full plate voltage on it when the amplifier is turned on. (Distilled water is supposed to be a good insulator, but don't bet your life on it.) A 3- to 5-gallon plastic bucket can be used to hold the pump and water for the complete system.

At K6HCP's successful moonbounce station, the 2C39's in the 300- to 600-MHz doubler, 600- to 1296- MHz double-mixer, two-tube 1296-MHz driver* and six 3CX-100A5's in the UPX-4 ring amplifier are all water cooled. After many hours of echo tests and moonbounce schedules, not a single tube has been replaced or even suspected of having low output. Best of all, for weak signal detection with a full kilowatt input, nothing but the purple glow of stressed 3CX100A5 ceramic disturbs the silence of the shack.

* Peter Laakmann, WB6IOW, "Cavity Amplifier for 1296 Mc," *QST*, January, 1968, p. 17.

Mike Staal, K6MYC

a modular fm communications receiver

Project report on a functional design using off-the-shelf modules

David M. Stahley, K8AUH, 6401 Brookside Drive, Cleveland, Ohio 44144

Among my interests in amateur radio is fm technology. About a year ago I outlined a project for myself to design and build a functional receiver with certain features needed to pursue my particular interests. I wasn't interested in a deluxe, high-performance commercial design, nor was I interested in a lot of embroidery. I wanted an inexpensive receiver for continuous mobile or base-station monitoring. The features of this fm receiver were to include:

1. Noise-operated squelch
2. Carrier relay
3. S-meter option
4. Minimum power consumption

I also wanted to use off-the-shelf components to minimize construction difficulties and to keep the cost within reasonable bounds.

The receiver described in this article fulfills all these requirements. It is, of course, a compromise between good engineering practice, state of the art and practical home construction. Some of its features are readily adaptable to your equipment, and for this reason I've included discussion material at each level. This isn't necessarily a construction article, but rather a report on the outcome of the project.

rf deck front end

The front end (**fig. 1**) was obtained from Plectron. It was available in both low band (30-54 MHz) and high band (144-174 MHz) models; the costs were

$10.00 and $13.00 respectively, less crystals. The crystals will depend upon your particular requirements and location. Another approach would be to use any of the many available solid-state converters. A particularly good example are those produced by Vanguard; they will "custom tailor" a converter (rf deck in our case) for $2.00 over the cost of the basic converter. This price includes the crystal. Additionally, units with dualgate mosfets are available, which are excellent. Should you elect this route, be sure to specify the desired input (monitoring) frequency and 10.7 MHz output. It seems logical that some cost reduction might be realized if the deck is ordered less case and plugs.

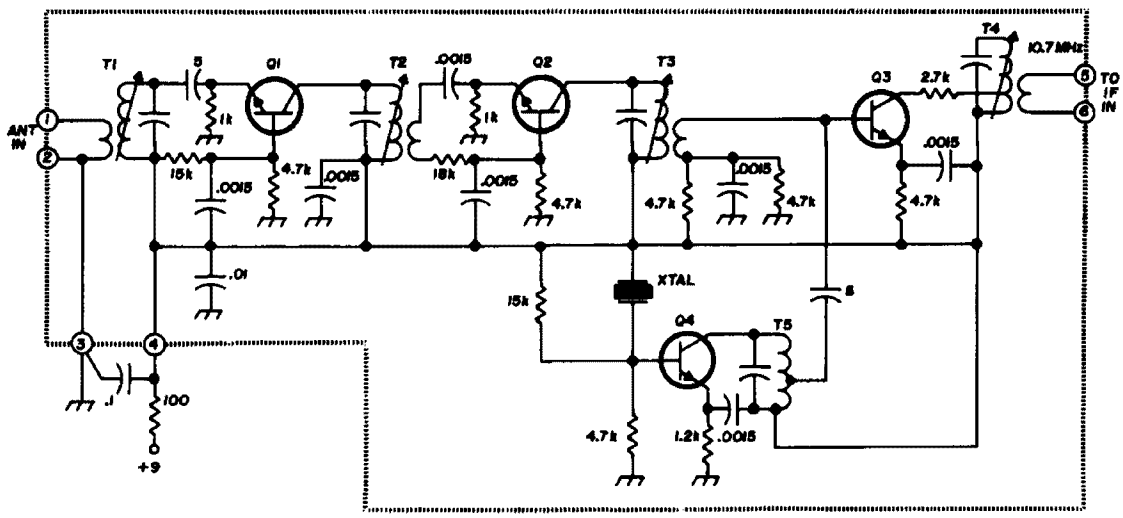


fig. 1. The Plectron low-band rf deck.

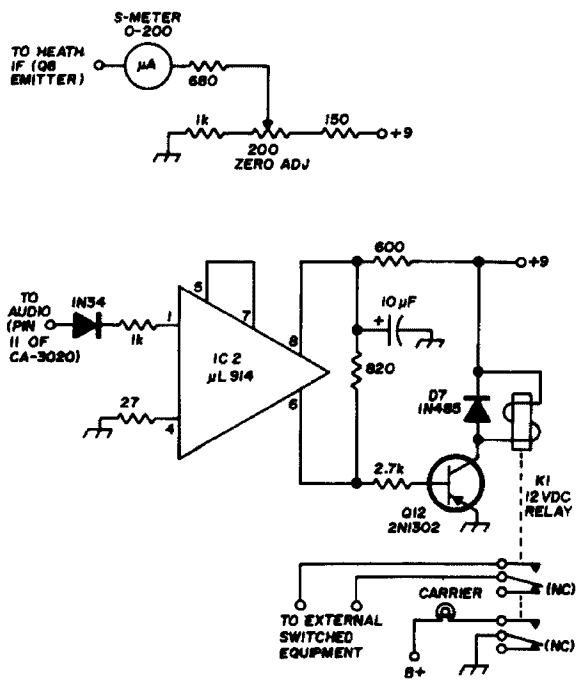
squelch

The noise-activated squelch (**fig. 3**) is simple but effective. The squelch transistor is a high-gain type similar to the Motorola HEP 54. Because of some poor experiences with noise squelch circuits in the past, I made several tests to ensure acceptable performance. Specifically, this circuit does not squelch off in the presence of high modulating frequencies such as are found in some selective-calling and remote-signalling systems. Neither is it affected by the mark (2125 Hz) and space (2975 Hz) tones used with audio frequency shift keying (afsk) radio teletype. Finally, because of its characteristics, mixed wide- and narrow-band transmissions cause no problem.

carrier relay

The optional carrier relay (**fig. 2**) has been used to actuate a tape recorder for

fig. 2. Optional s-meter and carrier relay circuit.



logging and recording. It's also very handy for operating a carrier indicator lamp. The usefulness of such a lamp becomes apparent if more than one receiver is operating in the same room. Any 12-Vdc relay can be used, provided it will pull in at 30-40 mA and drop out at 10-15 mA. The one used here is a Potter & Brumfield KM11D, which is a dpdt type approximately one cubic inch in volume.

rf and i-f sections.

This modification would resolve the problem but add considerably to the cost and complexity, which would move beyond the limitations noted earlier. I've shown how to add an s-meter to the unit (**fig. 2**); however, in this type of service it was found to be more window dressing than necessity. Any other compatible i-f strip could be substituted.

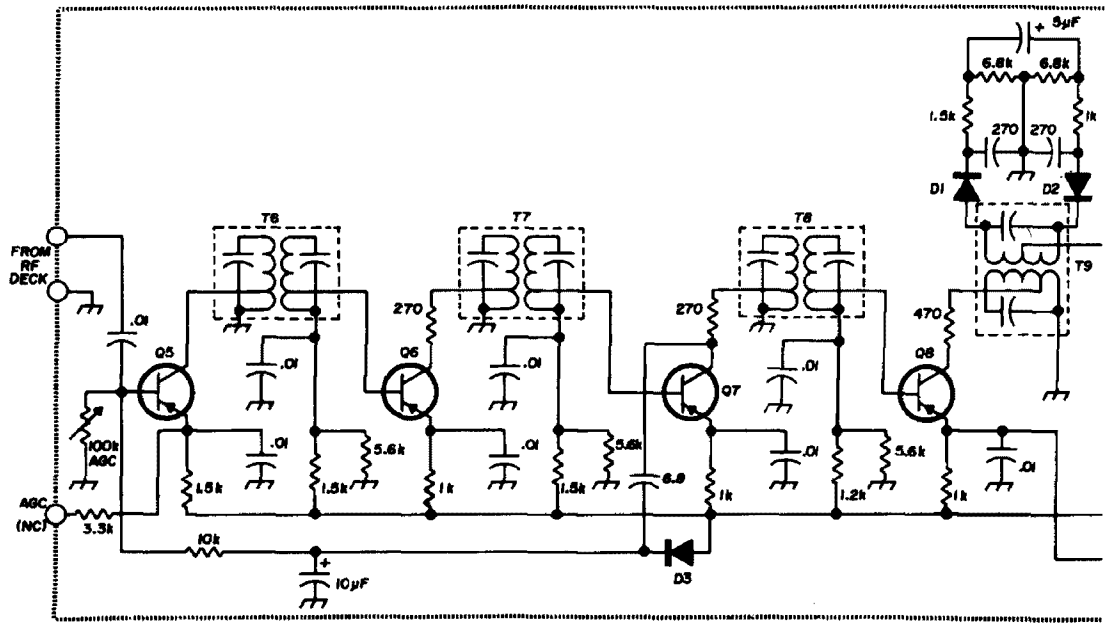


fig. 3. The Heath 10.7-MHz i-f strip (dashed box), squelch and IC audio amplifier.

i-f strip

The i-f strip is shown in the dashed box of **fig. 3**. It was one I found in the Heath AJ-43D receiver.[1] It was available from Heath complete and ready to operate for about $17.00. This choice results in an i-f of 10.7 MHz. Benefits include availability, commercial development, compact construction, low cost and relative freedom from drift. Disadvantages are relative low audio recovery and susceptibility to adjacent-channel interference. The first is overcome by additional gain in the audio stages. The second, where it might be a problem, can be reduced by addition of a ceramic filter and additional amplification by using an RCA CA3028 between the

audio power amplifier

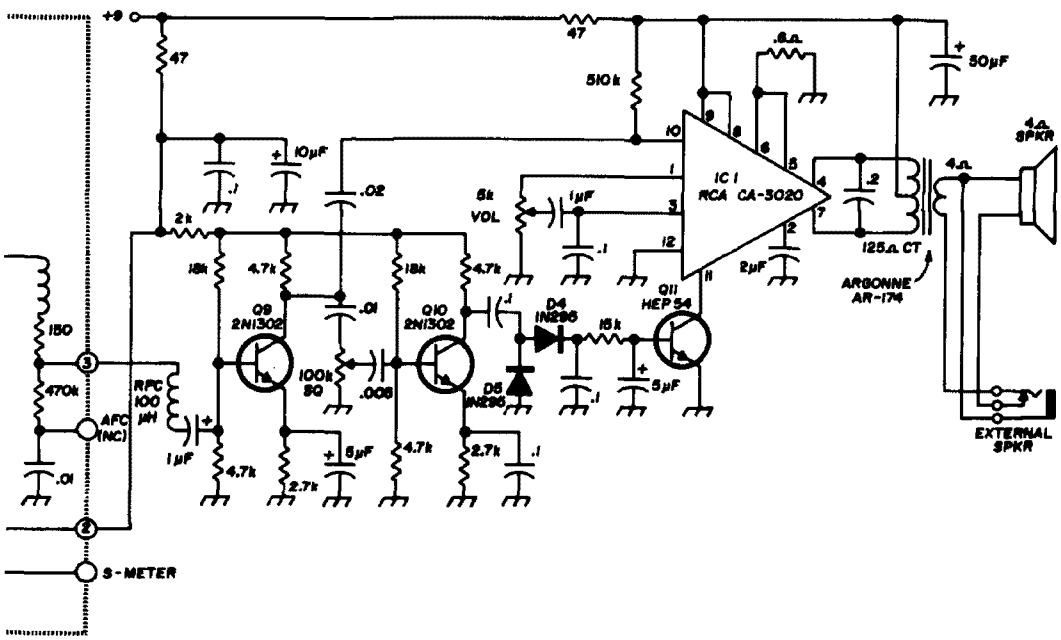
The audio output section (**fig. 3**) uses the RCA CA3020 integrated circuit. This is a high gain, self-contained audio amplifier capable of 500 mW output, which is quite adequate for indoor use. For mobile service, a better choice would be the CA3020A, which is rated at one-watt output. This should provide sufficient power to overcome most wind and traffic noise. The audio system frequency response has been limited to 300 to 3000 Hz to facilitate communications and reduce operator fatigue due to hiss and noise. A further extension of this would be to use a Jensen type PC4V3 four-inch speaker. The response of this speaker is similarly restricted.

power supplies

The power supply is completely open to choice. Four are shown in **fig. 4.** A nine-volt battery will work nicely, as will an ac supply consisting of a 6.4-Vac filament transformer and diode bridge. For mobile service a zener regulator or zener-transistor regulator would be in order to help tame the shifty voltages. Further, a small blocking diode is a cheap form of insurance against accidentally reversed battery polarities. The ultimate would be to include a rechargeable battery inside the case, with a regulated ac supply. Then you could unplug from the base ac mains, move to the car, change antennas, plug into the cigar lighter, start up and drive away—all without missing a word.

The problem of both ac and dc operation can be greatly simplified by using two power cords terminated with 4-pin Jones female plugs. This eliminates the need for changeover switching. One cord is made using pins 3 and 4 for 110 Vac. The other cord would be for mobile use, connecting pin 1 to +12 Vdc and pin 2 to vehicle ground. Don't forget the blocking diode. The battery can be kept charged while on ac power by means of a resistor and diode combination.

The next generation will probably have IC's in the rf and i-f sections. This hasn't been done as yet, because it's difficult to find commercially available i-f and detector transformers that match the available IC's. I'm interested in using ceramic i-f and discriminator units, such as are available from Murata and Clevite. However, so far everything in this whole area seems to be either wrong impedance, too wide- or too narrow-band, too expensive or generally not available.



Another nice item would be an IC suitable for an untuned crystal oscillator (10.245 MHz) and mixer unit for dual conversion. Also, a linear IC with an fet

materials and suppliers

1. Heath Company, Benton Harbor, Michigan 49022. Order: 10.7 MHz solid-state i-f strip P/O Heath Model AJ-43 or AJ-43D Part number 100-M521
2. Plectron Corporation, Overton, Nebraska 68863. Order: high-band rf board number 3185 ($12.00 ea.) low-band rf board number 3283 ($9.00 ea.); crystals for Plectron Boards HC-25, holders non-oven type, 0.005% third-overtone, 32 pF load ($8.00 ea.)
3. Vanguard Electronics Labs, 16-23 Jamaica Avenue, Hollis, New York 11423

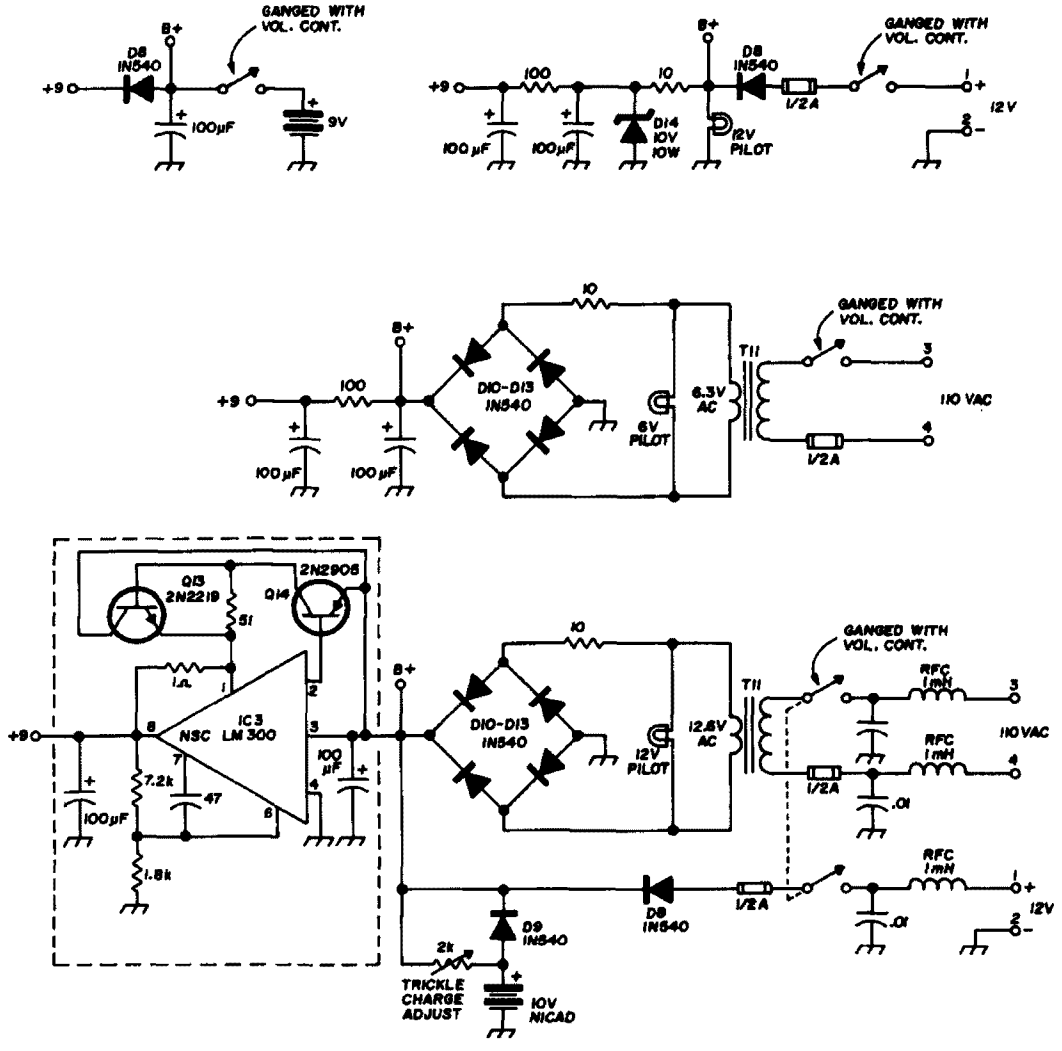


fig. 4. Power supplies for the modular fm receiver.

mixer section for the first converter/first i-f amplifier would be desirable. National Semiconductor has just released a complete squelch and agc unit, which would shrink the audio sections still further. Then there are LSI (Large Scale Integration) techniques, which could mean practically the whole receiver pasted on the back of the speaker; next year's receiver should be a lot smaller, a better performer and quite a bit cheaper and easier to build.

ham radio

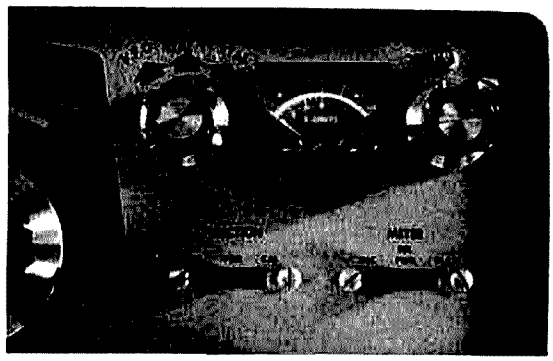
crystal control for the HW-100

Crystal control is an option not offered by most transceiver manufacturers. Granted, most vfo's are stable and accurate enough not to require the steadiness of a rock, but there are times when crystal control is a definite advantage. Mobile operation is enjoyed by a lot of amateurs, and here a stable, known frequency is most welcome. Many of my friends operate on a single frequency most of the time; there's someone monitoring the frequency nearly 24 hours a day, and this can be reassuring to a lone mobileer. Crystal control is also useful for net operation.

The crystal oscillator shown in **fig. 1** provides more drive than necessary and must be padded down with the 45-pF trimmer. The circuit is very inexpensive to build—less than the cost of the crystal itself. A positive supply voltage is picked off the zener diode that biases the vfo amplifier tube. With the switching circuit shown, it is possible to select: vfo-controlled transmit **and** receiver; crystal-controlled transmit and vfo receiver; or, crystal-controlled transmit **and** receive.

When ordering the crystal for this circuit, specify that it will be used with a 32-pF load. The vfo tunes from 5.5 to 5.0 MHz (low to high end of the band). Therefore, to operate crystal controlled on 3999 kHz for example, you need a crystal cut for 5001 kHz. The 12-pF trimmer is used to put the crystal exactly on frequency. If you plan to operate near a band edge, it would probably be a good idea to spend the extra money for a commercial-grade crystal.

The crystal oscillator is built up on a terminal-strip board and bolted to the back of the vfo chassis as shown in the photo. An insulating strip is placed under the board to keep it insulated from the chassis. Drill a hole beside the 6AU6 to accommodate an insulated feedthrough bushing: the 45-pF trimmer is mounted between the circuit board and the feedthrough insulator. Additional holes are drilled through the top of the vfo chassis to accommodate the wires needed to operate the crystal circuit and incremental tuning unit.



Front panel of the modified HW-100 showing the crystal/vfo control switch.

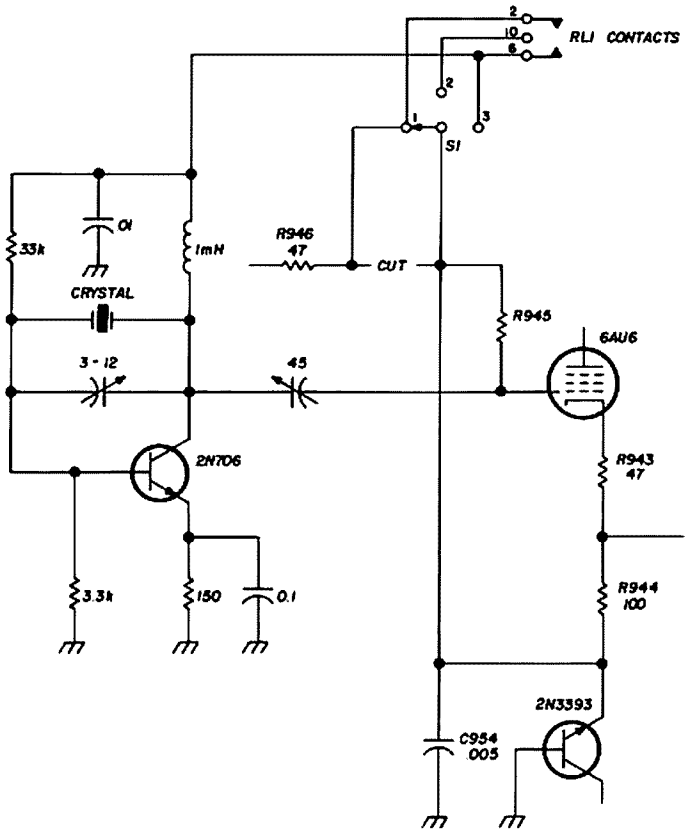
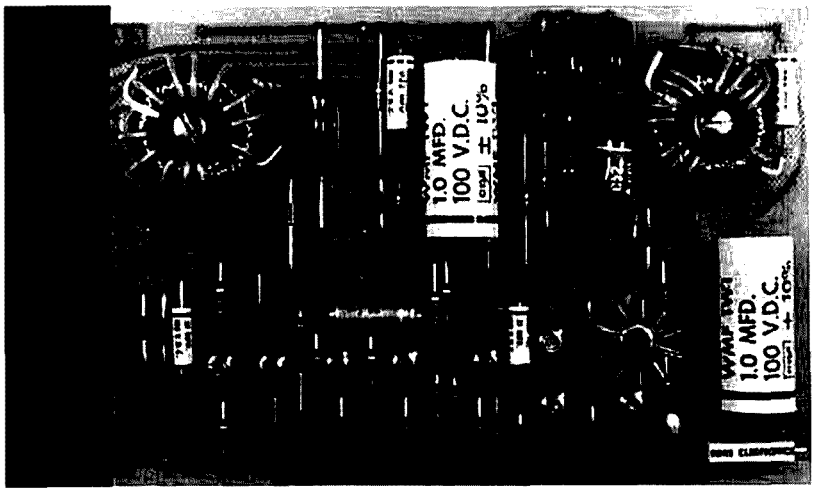


fig. 1. Crystal control for the Heathkit HW-100. In switch position 1, the vfo controls both transmit and receive; in position 2, crystal control on transmit, vfo control on receive; position 3, crystal control on both transmit and receive.

The 47-ohm resistor is disconnected from terminal strip FF (pictorial 7-2 in the HW-100 manual). A wire is soldered to this resistor and passed through the top of the chassis to terminal 2 of the rotary vfo/crystal switch: a wire soldered to terminal 1 of terminal strip FF is connected to the common terminal of the vfo/crystal switch: a wire coming from terminal 3 of the switch is connected to the B+ tie point of the crystal oscillator. Three more wires are connected to the switch as shown in the schematic and connected to terminals on relay RL1.

Bill McCracken, K1GUU



why not try RTTY?

Circuits for sending and receiving radio teletype—plus some station-improvement ideas

E. C. Sherrill, K6JFP, 4745 49th Street, San Diego, California 92115

While reading amateur radio publications I've noticed a very few simple, easy-to-follow construction articles on RTTY. To help fill the void I'd like to present some circuits that have added a great deal to the operating convenience of my station. Solid-state devices are used exclusively except for power supply transformers. For those who would like to learn more about RTTY theory and design, some references are included at the end of the article.[1, 2, 3]

basic equipment

The most expensive item in an RTTY station (excluding transmitter, receiver and antenna) is the teleprinter. As with most equipment, these machines can be obtained by judicious horse trading. I've seen some advertised for as little as thirty dollars. Rebuilt ones cost up to $150 if you want to go first class.

If you have a teleprinter, all you need to receive RTTY is a terminal unit (TU). The TU accepts tones representing code characters from the receiver, discriminates between the tones, and changes them into

positive- and negative-going pulses. These are sent through limiter and flip-flop stages and finally to a keyer stage. The keyer operates a relay in the teleprinter.

To transmit you'll need a frequency-shift keyer (fsk). This is a simple capacitance-tuned LC circuit that shifts the transmitter vfo output (usually 850 Hz) to form the code characters. The remaining equipment consists of two simple power supplies. Added refinements include an oscilloscope to monitor keying waveform and an audio frequency shift keyer (afsk).

The TU printed circuit board is available at nominal cost.* It is a heavy-duty fiberglass board, and information on parts layout is included. (Mine came equipped with all edge connectors and transistor sockets.) You can get the complete unit with all components, wired and ready to play, for about sixty dollars. Either package comes with complete and extensive technical data.

The TU is the basis of the entire system (**fig. 1**). I modified the original Cashion circuit to accommodate some switching circuits for added operating convenience, and I added the power supplies and fsk board. A schematic is also included for an afsk board (**fig. 5**) for those who might be interested.

power supplies

These are diagrammed in **fig. 2**. Jacks J3, J4 and J5 must be insulated from the TU panel. A transmit/receive switch (S5A) gives control of the transmitter from the RTTY position. (See also **fig. 3** which shows connections between TU board and transmitter.) Switch S5B, which is one-half of S5, disconnects the TU keying circuits during transmission but lets you monitor the tones with an oscilloscope. (see also **fig. 1.**)

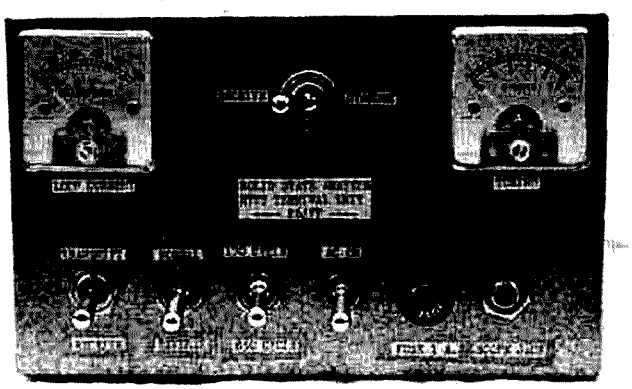
the TU board

Most RTTY terminal units have provisions for normal/reverse copy. Occasionally you'll find a station using reverse transmission, or you just might have tuned him in on the wrong sideband. Switch S2 (**fig. 1**) will give the correct copy. Also I added a narrow/wide control; this allows you to choose between 850- or 170-Hz shifts, which are most commonly used.



The fsk with accessory cables and a dummy plug for the TD.

Front panel of the solid-state TU.



My TU has a three-circuit microphone jack on the front panel. This is used to monitor signals with an oscilloscope, either during transmit or receive. Signals vary in amplitude, of course, and the S-meter is handy; but a scope is much better, especially during net procedures.

In the TU panel photo the switches are, from left to right, transmit/receive switch S5; normal/reverse switch S2; narrow/wide

* Cashion Electronics, P. O. Box 7307, Phoenix, Arizona 85011. Order a "two-tone limiterless TU no. 6." Price is about $10. A phenolic board, without holes or plating is available for $5.

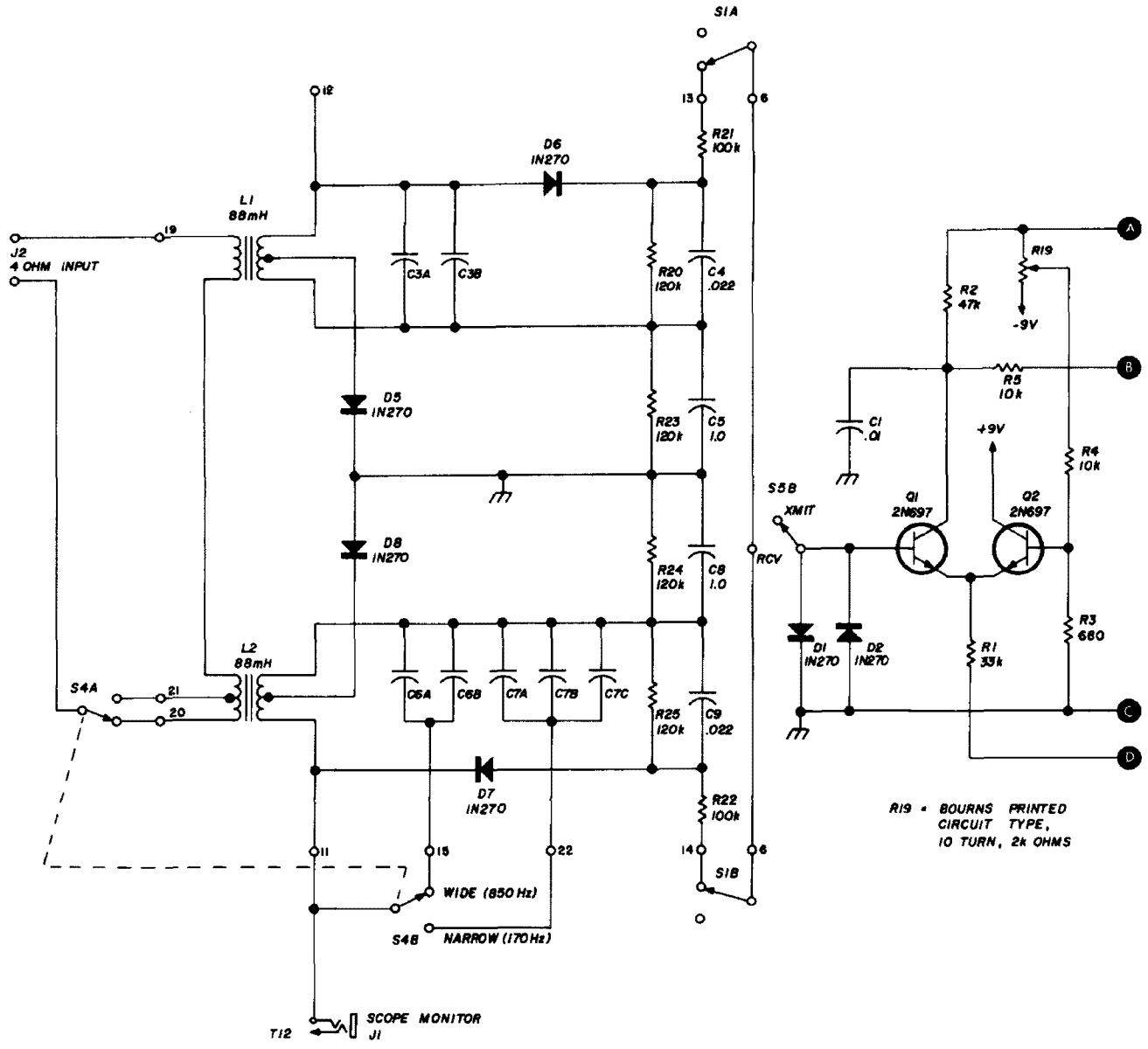


fig. 1. Schematic of the tuning-unit board. This is the basic unit; the text explains how accessories were added.

switch S4; and ac switch S3. Next to these are the fuse holder for F1 and oscilloscope jack J1. At the top are the 0-100 mA dc loop meter, receive/standby switch S1, and the tuning meter.

For transistors Q1-Q8 I used 2N697's, but other more inexpensive ones are now available. I used an RCA 2N3440 for Q9; again, these aren't absolutely the final word. Other types will work equally well. Be sure to use a heat sink with Q9.

the fsk unit

This circuit is shown in **fig. 4.** Parts are mounted on a PC board measuring about two by three inches. **Fig. 3** shows how the fsk output is brought into the TU board. The fsk is connected to the vfo in the transmitter by a single loop of wire wrapped around the vfo-tube cathode.

construction notes

The photos show the general layout of parts and the interconnecting cabling. Recognized techniques and precautions should be used when assembling these solid-state components. I shall emphasize again that transistor Q9 on the TU board **must** be installed with a heat sink.

Table 2 gives the connections to the TU

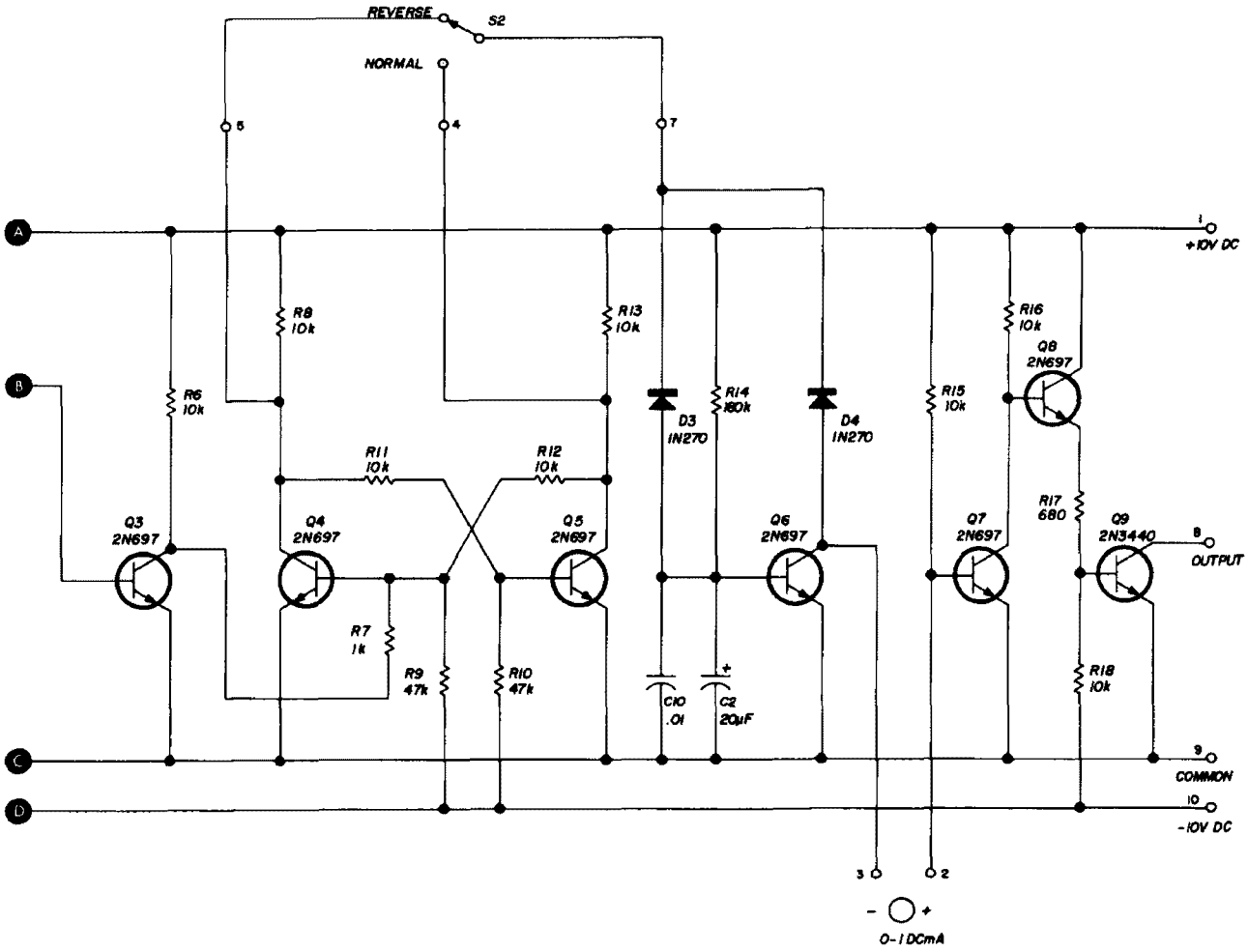


table 1. Parts list for the power supplies, fsk, switching circuits and accessories. (Note: Does not cover the TU board. See fig. 1.)

Jacks

J1 & J6, 3-circuit microphone jacks
J2, J3, J4, J5, J7 single-circuit phone jacks

Switches

S1 dpdt mark/space—standby
S2 spdt normal/reverse
S3 spst ac switch
S4 dpdt narrow/wide shift
S5 dpdt transmit-receive

Meters

1 ea 0-100 dc mA Lafayette Miniature
1 ea 0-1 dc mA or VU Lafayette Miniature

Cabinet

1 ea 5 x 7 x 9 inches Cal Chassis Company LTC 463

Transformers

1 ea Triad R-30 or equivalent
1 ea 120 V primary, 15- to 24-V secondary, 100 mA, Triad F-45X or equivalent

Fuses

F1 1A, 150 V

tu terminals	functions
1	+10 Vdc from regulated power supply
2	plus terminal of 0-1 dc mA VU tuning meter
3	minus terminal of 0-1 dc mA tuning meter
4,5,6	normal-reverse switch S2
6	To switch S5B mark space tones input to board
8	connect to RY-KBD-TD insulated jacks via loop supply
9	common ground
10	minus 20 Vdc from regulated power supply
11, 15, 22	vertical scope input, switch S4B or tuning bulb wide/narrow tuned circuit
12	horizontal scope input or tuning bulb
16	nc
17	nc
18	nc
19	from receiver output, 4-ohm input (J2)
20, 21	switch S4A. wide/narrow shift tapped coils

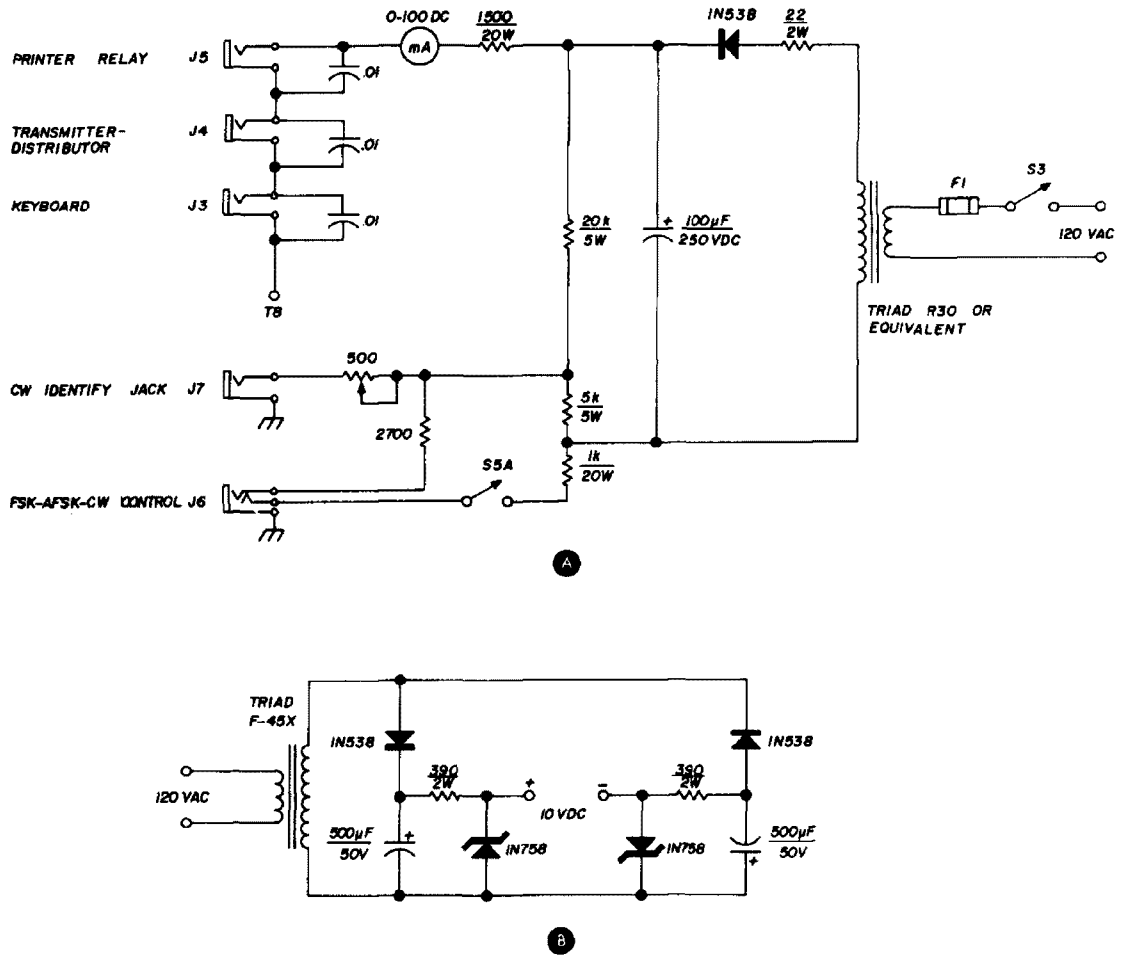
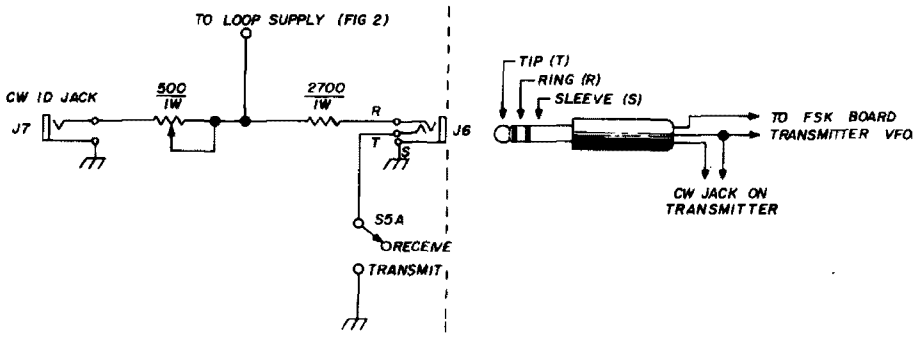


fig. 2. The loop supply (A) ties into the TU board at terminal 8. The low-voltage supply is connected to terminals 1 and 10. See fig. 1.

board. These jacks are shown in **fig. 1.** The power transformers and TU printed circuit board are mounted above chassis; all other parts are mounted below chassis. The front panel of my TU is secured by the switches, fuse holder and jacks. I used a Triad R-30 transformer for the 120-volt supply and a Triad F-45X for the low-voltage supply, although any equivalents will work satisfactorily.

After everything is assembled, check all connections and wiring, then you're ready to insert the plugs. Turn on the power and check the 100-mA meter. It should indicate

fig. 3. Fsk connections between the terminal unit and transmitter.

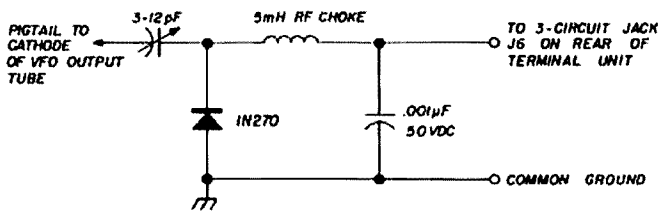


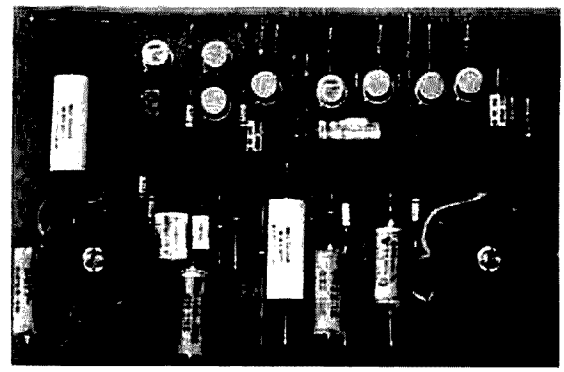
between 55 and 70 mA. If not, something is wrong, and you should recheck the circuits.

If the meter reads okay, turn on the audio, and then watch the loop meter and tuning meter. If signals are tuned in properly, the teleprinter will print out correctly. If it doesn't, check the beat frequency oscillator sideband in use, or the normal/reverse switch.

If you don't have the proper test equipment to tune the fsk unit, load the transmitter on a dummy load and vary the 12-pF capacitor. Try to match the received mark and space signals on your oscilloscope. When the tones are nearly correct, put your transmitter on the air and call a station. Your mark/space signals might not be just right, but he'll receive **something.** Then the real spirit of amateur radio will prevail!

fig. 4. Tho fsk circuit; this provides 850-Hz deviation for RTTY characters.





The Cashion TU-6 board as assembled by the author.

some parting thoughts

I am quite pleased with the circuits described. If you try them and have any real problems, send your questions and a self-addressed, stamped envelope. I'll do my best to help you. The most important watchword when working with these units is: **get the polarity right the first time.**

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2. Durwood J. Tucker, W5VU, "RTTY from A to Z," CQ, August, 1964, through July, 1966.
3. Irvin M. Hoff, K8DKC, series of RTTY articles in QST during 1965: January, p. 14; February, p. 29; March, p. 24; April, p. 44; May, p. 16; June, p. 32; August, p. 27; October, p. 21.

ham radio

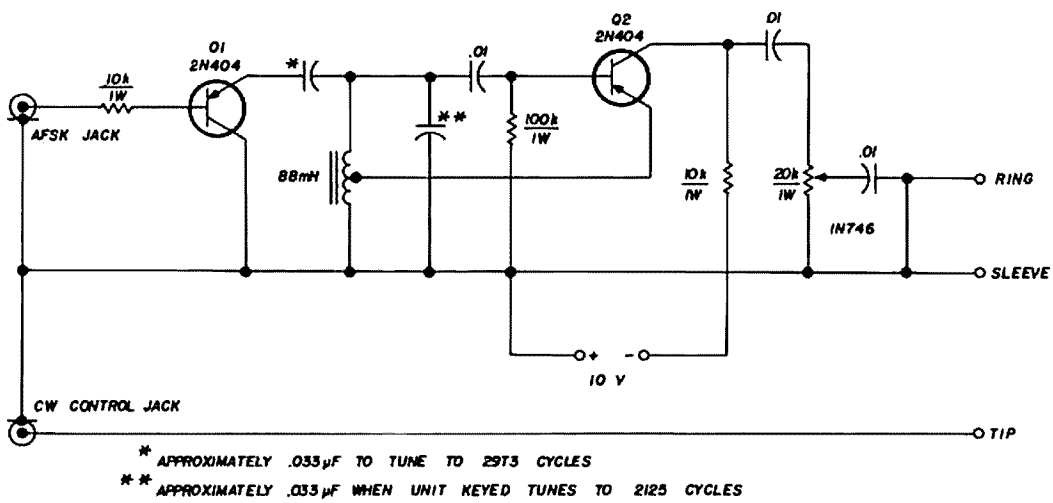


fig. 5. The afsk unit. Directly replaces the fsk if you wish. Remove the cable from the fsk and plug into the afsk. Output plug of the afsk (PL-68) plugs into an a-m or fm transmitter.

grounding and wiring

Jim Ashe, W1EZT

When you are trying to make a circuit that you've built from a schematic work properly one of the problems that may plague you is the grounding bug. This has probably caused more instant insanity than any other problem facing the electronics experimenter. As with any worthwhile project, there's a right way and a wrong way to do things. I'd like to pass along the results of some of my experiences with ground connections and wiring. If you're interested in exterminating the grounding bug on your next construction project, read on. It may save you much time and trouble getting the circuit to play.

grounding and wiring principles

Successful grounding in any electronic circuit depends on how dc voltage and current are controlled by wiring placement. Voltage can do strange things to your circuit. Reason: voltage is a workhorse which, if not treated with respect in circuit wiring, can be most unforgiving. Depending on the circuit, the negative pole of voltage may be grounded at a common point on the chassis. Or it may be left floating, as in a transformerless circuit that works directly from the ac input source.

Current, in contrast to voltage, doesn't terminate at some point on the chassis, but makes a complete pass around the circuit. Anything that interferes with the current loop (stray capacity and inductance) reduces circuit efficiency and produces an instability problem.

In the following discussion I shall explode some myths about grounding that have hung around since the early days of radio. If you consider what I have to say, you should have a better understanding of how voltage and current behave in electronic circuits. Then you can construct your equipment with these thoughts in mind and be reasonably sure that luck, at least, won't be one of the independent variables that influences hardware performance.

about voltage

Digressing for a moment, voltage is electrical pressure between two points. It's measured by the indication on a meter connected between the two points. Commonly, one of these points will be at ground potential. One definition of ground is "a circuit point that may be earthed without upsetting the circuit." This definition was valid for circuits years ago, but for today's modern circuits it is an oversimplification.

The circuit designer has a choice as to whether a ground will be included, and if so, which part of the circuit will be at common ground potential. Usually a large piece of metal, such as a chassis, is chosen as the place where you attach one lead of your meter to measure dc voltages (**fig. 1**). The circuit is usually designed so that signals look good when viewed from this reference point.

The manner in which voltages are controlled in circuits can mean the difference between success and failure when you first turn on that new piece of equipment. The schematic is an idealized road map. It does not show the many extraneous resistances, capacitances and inductances found in the hardware. If all these elements were included in the schematic it would be quite messy, indeed. Good wiring practice and lead dress will minimize these problems, which are the grounding bugs I mentioned earlier.

into the input circuit. Then you have feedback, and an unstable system.

about circuits

Modern electronic circuits seem to be harder to work with than those of a few years ago. Active devices have more gain, which means more sensitivity to unwanted feedback. The higher gain is often accompanied by response to much higher frequencies. This means smaller unwanted coupling inductances and capacitances can return enough signal from output to input

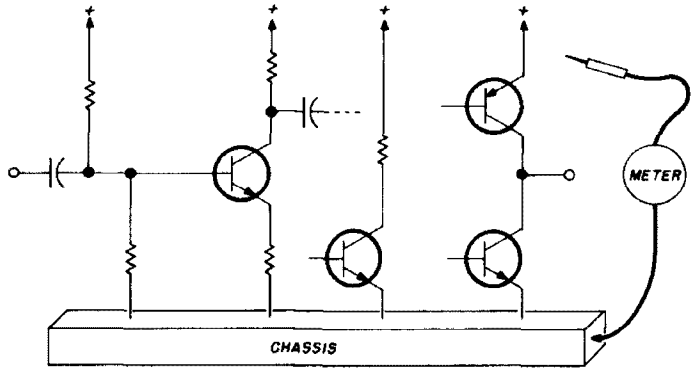


fig. 1. When you think of a circuit ground, you usually think of something like this. Everything in the circuit refers to the one solid piece of metal, and test meters are attached to this reference point.

about current

Current can be thought of as a kind of fluid, but flowing in a solid medium rather than in a pipe. If you want to measure current, you insert a measuring device into the stream, or you may insert a resistor into the stream and measure the voltage developed across it.

The water pipe analogy for current flow is misleading, at least when you're trying to understand what happens to current flow in an electronic circuit. Eventually water in a pipe runs down a drain somewhere. But current makes a complete pass around an electronic circuit, as in **fig. 2**. There's no common point where current can be disposed of, as with voltage. Current pops up again somewhere else, and when you're building a circuit you must know where.

A current gone astray can cause unwanted feedback. **Apparent** grounding of an audio amplifier, for example, can fool you. If you choose just any handy tie point, some of the output current can find its way

to produce instability. Have you noticed construction projects lately containing RC circuits or ferrite beads in their input circuits? These are added, as in **fig. 3**, to control high-frequency response. Older circuits didn't need these provisions.

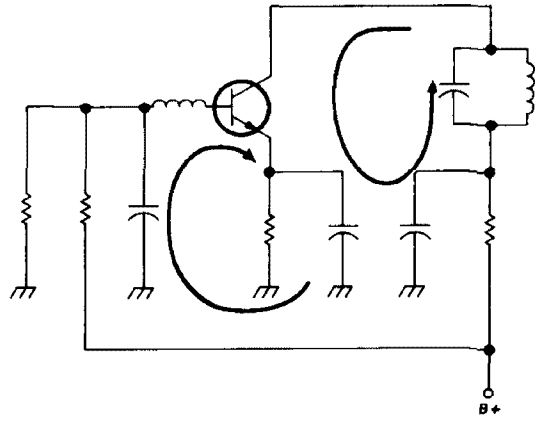


fig. 2. Input and output currents follow loop paths in this single i-f amplifier stage. Interfering with the current loops reduces efficiency and increases hazard of unwanted feedback.

A schematic contains two circuits: the dc supply circuit and the ac signal circuit. This fact is obscured in the schematic, because many components perform two jobs. The collector resistor in a transistor audio amplifier, for example, functions as a signal load resistor as well as a device to decrease the supply voltage to a lower value as seen at the collector. Let's trace these circuits and see what happens (**fig. 4**).

From the power supply positive terminal, the dc bias current flows along the supply line. Then it reaches a point where it can flow through the load resistor to ground. Going back, we find other routes: through the base bias network, and from base to emitter to ground. At the ground terminal these currents meet, and we trace them back to the power supply. This completes the current loop. It shows that the dc-supply circuit actually extends through the active elements in the circuit. Thus there's a lot more to the dc supply than two wires with the circuit between.

Now let's look at the signal circuit. We trace the current loop the input signal follows. The signal flows to the base, down to emitter and ground, and back to the input. The collector signal flows to the emitter, to the emitter bypass capacitor, then to ground, and finally back to the collector.

Here's an important point, sometimes easily missed: not **all** of the signal follows this route. No matter what resistance is placed in the collector circuit, some of the signal finds its way to the supply lines and to all the places to which they extend.

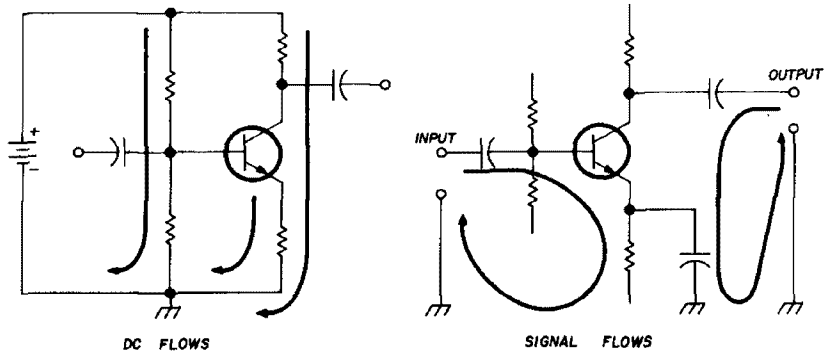


fig. 4. Signal and dc biasing paths share elements in this simple audio amplifier circuit.

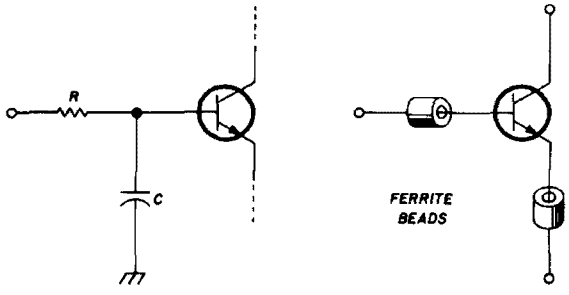


fig. 3. Two methods for reducing high-frequency response.

This is why some writers emphasize decoupling.

Now for some practical examples.

oscillators

In **fig. 5** I've drawn a simple oscillator circuit. It can't feed currents into the rest of the system, because it has no signal ground. It could be operated from a negative supply by returning the positive terminal to supply ground, with the negative terminal above ground; this is the commonbase oscillator. There's no magic about having the base connected to a large piece of metal, so this connection is omitted. All currents have free paths to follow correct routes, and the circuit works fine.

audio amplifiers

I once had an experience with an intercom amplifier that simply would not stabilize. It turned out to be caused by a speaker ground return to a handy tie point under the gain control. Let's see

what happens when such an amplifier is wired according to the single-point ground theory.

The circuit of **fig. 6** is one you might assemble from a printed-circuit amplifier kit. Why isn't it stable?

Heavy output current is provided to the speaker. One watt into eight ohms is 125 mA. The speaker return is connected to the

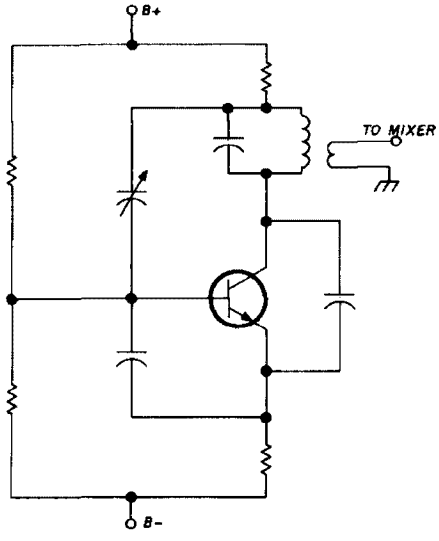


fig. 5. Here is a conventional common-base oscillator with no ground connection for feeding a simple rf mixer.

chassis. After passing through the speaker, the current flows all the way around to the common ground wire and finally back to the power amplifier.

Probably because of the way input connectors are constructed, and because there's no extra terminal on the PC board, the input current must follow a route through the amplifier, along part of the same wire to the chassis, and back to the input connector. This doesn't look too bad. But let's estimate the voltage produced by the output current across the input terminal.

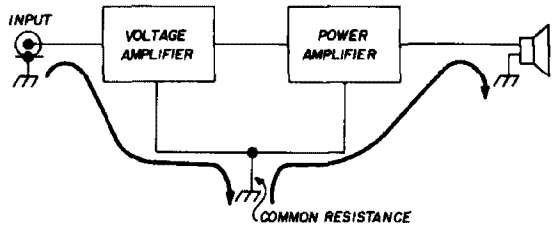
If the entire resistance of the common wire, including chassis connection, is 10 milliohms (and with a poor connection it could be more), the 125 mA audio current can generate 1.25 millivolts of unwanted current. If the phasing is correct, oscillations may result. At some frequency, the phase probably will be correct. The remedy is a small revision of the output circuit. This will allow the heavy output current to follow a **direct** route to the output stage.

rf amplifiers

The grounding bug can be responsible for rf feedback in rf stages. I once talked to a fellow with a well-shielded but unstable kW rf power amplifier. "How can this be unstable?" he asked. I don't know how it finally came out, but I saw something that appeared to be a prime suspect.

Rf current flowed from the two 813 anodes to the pi network. From there, it flowed into the antenna coax center conductor. The key here is that equal return current came **back out** of the connector and eventually found its way down through the two holes in which the tubes are placed. Finally, the current found a way to return to the tube cathodes. If the phasing is correct, current

fig. 6. This audio amplifier is unstable because of feedback through a resistance common to both the input and output.



coupled into the input circuit will cause oscillation.

This problem can't be resolved by shielding, because that's not where the trouble is. Some provision must be made to carry the return rf current **directly** to the cathodes. My recommendation, before building any rf power amplifier, is to draw a picture of the current route, including the coupling system. Include any direct paths that the current could follow. If you can't see the route, the current probably can't either, and some rerouting of return leads is a must to tame the circuit. This will yield an additional benefit of improved sensitivity (in a receiving system) and better performance.

ham radio

a top-loaded 80-meter vertical

This efficient antenna can be built for under twenty dollars— it's one answer to restricted space

George Cousins, VE1TG, Box 18, RR 2, Lower Sackville, Nova Scotia

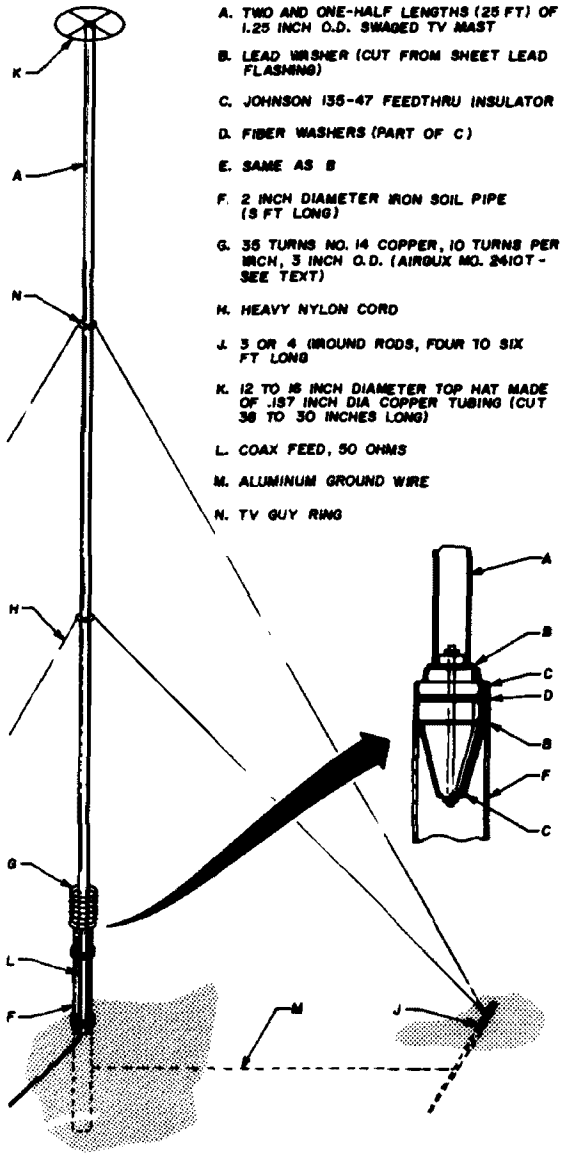
The search for an efficient, low-cost antenna for 80 meters often takes on strange forms, mostly because of the fairly considerable space necessary to get any sort of decent performance. Limited lot sizes often prevent erection of anything like a proper horizontal dipole, and even a long wire becomes a problem. However, a convenient answer to the problem may appear in the form of a vertical.

Many thousands of hams operate 80-meter mobile with small whips, using their car bodies as ground planes. Many attempts have been made to use these whips in home installations, with limited success. I believe the best answer is a happy compromise between whip and full-sized antenna.

The array described here is just such a compromise, but quite an efficient one, which should find ready appeal because of its low cost, small size and ease of installation. It's constructed of readily available materials, and can be built and installed in less than a day with no strain whatsoever. This antenna was designed and built by Wilbur (Bill) Hills,

VE1KK, a well-known 80-meter man here in the Maritime Provinces. Reference to **fig. 1** will provide clear construction details.

fig. 1. Construction of the low-cost vertical antenna for 80 meters.



construction

The radiating section is made from two and one-half ten-foot lengths of tv mast, approximately 1¼ inches in diameter. When purchasing the mast, pick up a couple of tv guy rings at the same time. The base support is made from a length of two-inch iron soil pipe (a five-foot length will be adequate). The pipe will be eventually buried in a hole where the antenna will be positioned, but don't dig the hole until you finish this article so you have an idea of just where to dig it.

Also remember that the soil pipe must be set truly vertical, or the antenna will have a drunken lean to one side or the other. To insulate the antenna from the mount, a Johnson number 135-47 feedthrough insulator is used. The insulator is inverted and its longest section is placed inside the pipe. The antenna is then placed on the insulator's base. To protect the porcelain from contact with the metal pipe, two washers cut from a sheet of lead flashing separate the insulator from the metal.

The heart of the antenna is the loading coil at its base and the capacitance hat at the top. Both are easy to construct. The coil is made from a section of commercial Air Dux number 2401T, or homemade using 35 turns of number 14 copper wire, wound ten turns per inch three inches long. Small plastic strips, similar to the commercial stock, should be cemented to the wire for support. Alternately, the coil can be wound on an insulated form.

The capacitance hat is 12 to 16 inches in diameter, using 3/16 diameter copper tubing or a close equivalent. If the outer ring is soldered to the radials, and the inner ends of the radials soldered to a metal ring clamp, the hat will be rigid and can then be clamped or even soldered to the top of the mast.

final assembly

Because the efficiency of a vertical antenna depends greatly upon its ground system, provision will have to be made to lay out as many ground radial wires as possible. The radials can be buried in the ground a few inches, although laying them on top of the earth will be all right. With this in mind, as well as remembering that the antenna must be guyed, proceed to lay out the position of the antenna and its guy stakes.

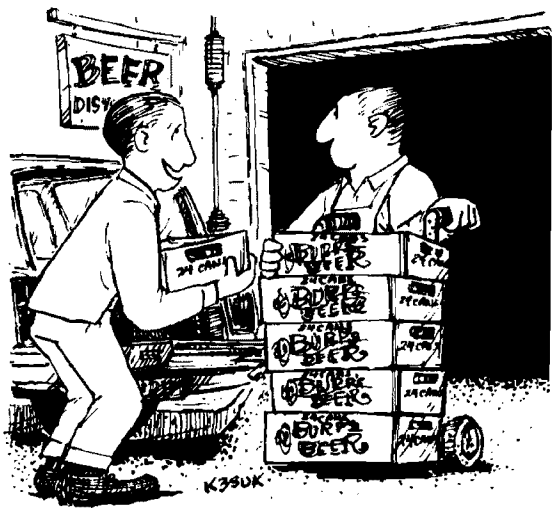
Dig the hole and set in the soil pipe. Next place the insulator, with the bottom lead washer, into the pipe. Slip the loading coil over the pipe and let it slide down onto the ground for the moment. The antenna element can now be assembled, and the guy rings installed. Because the element is very light weight, heavy nylon cord can be used for guys, and these should now be fastened to the rings. Two sets of three guys should be quite sufficient.

Before raising the mast, mark out the position of the guy stakes, and install them in a triangular pattern around the base. Ground rods a few feet long will be fine. Now the mast can be stood up, with its base near the soil pipe. When it's vertical, carefully lift it straight up and set it on top of the insulator. (Don't forget the lead washer.)

At this point I'll assume you have a friendly helper, and that he, or she, can run around and snub the guys while you hold the antenna upright. This will have to be done again later, to get the mast perfectly vertical, but at least it won't come down around your ears!

Now, slide the loading coil up the pipe and fasten its top end to the bottom of the mast. Slip a couple of small plastic strips between the coil and the pipe to hold it in place and prevent it from shorting against the metal. Connect the other end of the coil to the center conductor of the 52-ohm coaxial feedline, either RG 8/U or RG 58/U, depending upon the power you plan to run. The RG 58/U will be fine for up to a half kilowatt at least.

Bond the coax outer conductor to the junction of the ground radials, which can be made from copper or aluminum wire. Although they should be laid out radially, fanning out from the base in all directions, in practice this will probably be impossible. Follow the simple rule of "the more, the better," and lay them out in any pattern allowable. If the guy stakes are made of standard four- to six-foot ground rods, three of the radials could be also connected to these rods to give just a little bit better efficiency.



**"Oh yes . . .
I'll need a quart bottle
for the base insulator . . ."**

tune-up

Tune-up is simply a matter of matching the feedline to the vertical element. Use a clip-lead and tap the feedline along the loading coil until the best match (lowest swr) is obtained. It should be possible to obtain an swr under 1.5:1 for approximately a 300-kHz section of the band. However, like all shortened antennas, the swr won't be this low across the whole band, and some provision will have to be made for retuning if you make a radical change in operating frequency, such as shifting from the low CW section to the higher part of the phone allocation.

performance

Although the vertical antenna seems to be more useful for DX work because of its lower angle of radiation, this shortened version has good efficiency at the higher angles and has proved to be an excellent antenna for all-around work on this band. My thanks to my good friend VE1KK for passing along the design information and operational results that formed the basis for this article.

ham radio

cw selectivity with crystal bypassing

You can improve selectivity in your ssb transceiver with this simple filter— a no-holes modification scheme

John J. Schultz, W2EEY, 40 Rossie Street, Mystic, Connecticut 06355

If you are like most amateurs who own commercially built equipment, chances are you skip over modification articles no matter how good the circuit improvement might be. But if the modification can be added without drastic changes, then perhaps you might be interested.

This article describes just such a circuit. By using existing controls on your ssb transceiver or receiver you can enjoy the benefit of added selectivity for cw reception. The circuit is simple and effective. All you will need are a crystal, a switching diode, and a couple of resistors and capacitors. No holes to drill, no extensive digging into the existing equipment, and no worries about defacing panels or chassis. Although primarily intended for transistor or IC stages, this selectivity circuit will work with some sets using tubes.

adding cw selectivity

Adding a crystal filter in the i-f section of a transceiver or receiver, particularly to improve cw reception, often presents problems. The conventional approach is to break the signal connection between the i-f stages to insert the crystal filter. This requires some means of switching out the crystal filter when it's not desired. Also, unless the i-f strip has sufficient reserve gain, the insertion of the filter may cause a serious decrease in sensitivity.

The crystal bypass filter presented here was developed to overcome most of these problems. It's particularly applicable to ssb transceivers, which require better selectivity on cw than provided by the ssb filter. The ssb filter provides good skirt selectivity so all that's needed for good cw reception is a single crystal filter whose frequency is centered in the transceiver's filter passband. The crystal bypass filter and the ssb filter work in tandem.

This method won't give quite as good results as a six-pole, 400-Hz cw filter, but it will provide better cw reception under crowded band conditions than is possible by using the ssb filter alone.

The only other simpler way to provide cw selectivity would be with an outboard audio filter. The one great disadvantage

of this method is that the agc can't be disabled in any present-day transceiver, so strong signals entering the ssb filter still control agc action, although the audio filter makes them inaudible.

crystal bypassing

The simple idea of crystal bypassing is shown in **fig. 1A**. The emitter swamping resistor bypass capacitor provides no selectivity and, in fact, is chosen so the emitter is grounded for ac through an extremely small reactance at frequencies the stage is expected to amplify. If the swamping resistor were not bypassed, increased signal input would produce bias across the swamping resistor opposite to the forward emitter-base bias, and stage gain would decrease considerably. This degenerative effect is used sometimes to stabilize a stage, and also is the basis for many compressor circuits.

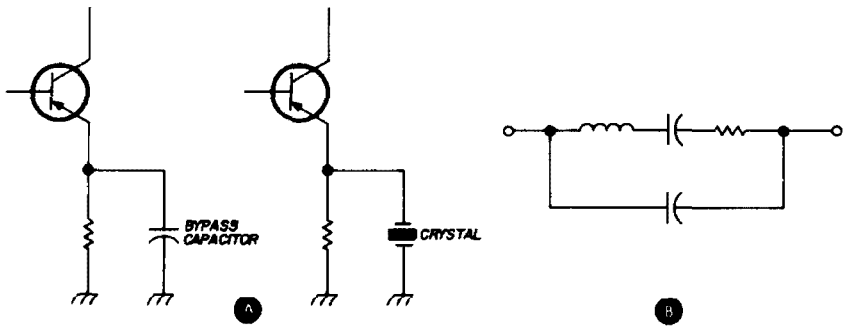


fig. 1. Capacitance and crystal bypassing of a common-emitter stage (A). Equivalent electrical circuit of the crystal is shown in B.

equivalent circuit

The equivalent electrical circuit of a crystal is shown in **fig. 1B**. It has two resonance modes, one formed by the parallel LC combination (high impedance) and one formed by the series LC combination (low impedance). The two resonant frequencies are very close together. While the series-resonant condition can't be externally influenced, the parallel-resonant condition is easily affected by wiring and crystal holder capacitance.

If the emitter bypass capacitor is replaced by a crystal, the series-resonant mode of the crystal will ground the emitter for ac at only one frequency. At all other frequencies, the stage will be highly degenerative and operate at reduced gain. The degree to which the crystal grounds the emitter will determine the loss introduced by the crystal. With crystals of high Q, which have very low equivalent resistance at their series-resonant frequency, the loss is small. In fact, it's conceivable that if an i-f stage were inadequately bypassed to start with, the crystal bypass might cause the stage gain to increase at the crystal frequency. On the other hand, if i-f stage gain is independent of the bypass capacitor, the crystal bypass wouldn't provide significant selectivity.

practical application

The bypass capacitor should first be lifted from ground to determine whether the crystal bypass method might be effective with a particular amplifier stage. The reduction in gain will indicate the broadness of the crystal response. If there is little gain reduction, signals outside the crystal passband will come through at about the same level as those in the passband. If a considerable reduction in gain occurs and the bypass capacitor is fairly large (0.1 μF or so), indicating the circuit is of low impedance, crystal bypassing will work well.

With conventional grounded-emitter transistor circuits, crystal bypassing will be most effective. However, other bypass capacitors can be tried until the most effective gain control action is found.

Fig. 2A shows how crystal bypassing can be added to an existing i-f amplifier. The crystal is inserted between the existing emitter bypass capacitor and ground. The capacitor should be increased to 0.1 μF if it's not this value already. An spst switch shorts out the crystal for normal operation. Although the coax cable (**fig. 2A**) should be as short as possible, it's capacitance doesn't affect the circuit greatly. Cable capacitance primarily affects the parallel resonant mode of the crystal, since the cable is shunted across the crystal.

Fig. 2B, a variation of the circuit in **fig. 2A,** was tried on an SB-34 transceiver. Diode switching eliminates an rf-cable run from the emitter and takes advantage of a switch position already provided on the front panel. A similar scheme should be possible with many other transceivers. Switch S1 transfers the meter from a 1-ohm shunt in the amplifier plate lead to an rf output-level circuit. The diode is connected to the switch arm that goes between ground and the plate-voltage line; thus forward bias is either applied to or removed from the diode. When forward-biased, the diode shorts the crystal for normal reception. The 1-megohm series resistor limits the diode forward current to about 500 μA.

The considerations outlined above also apply to crystal bypassing in vacuum-tube i-f stages. However, the problem here is that voltage levels are usually high enough to damage most crystals. The series capacitor between crystal and emitter in **fig. 2A** will act as a dc blocking capacitor when inserted between the i-f tube plate and crystal. If the capacitor is large enough, the crystal frequency won't be affected. A value between 0.1 and 0.25 μF should be about right.

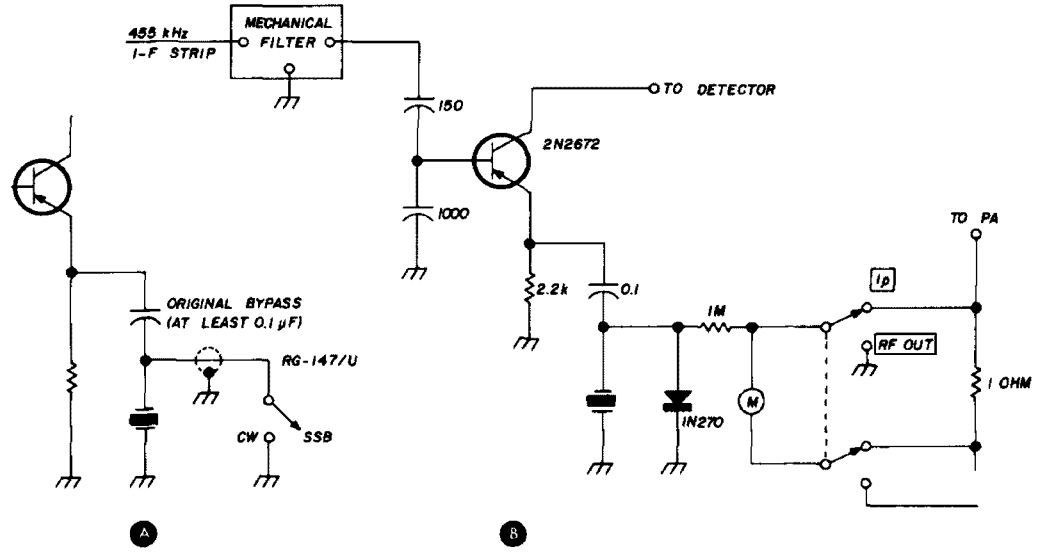


fig. 2. Simple switch-controlled bypass circuit (A). A diode-switched crystal bypass for an SB-34 transceiver is shown in B.

a final word

If you would like to add better cw selectivity to your ssb equipment, by all means consider the crystal bypassing scheme presented here. The circuit won't work in all ssb gear, but you can determine this easily enough as described previously. Careful examination of the front-panel control functions on most equipment should reveal one that will perform the auxiliary job of a crystal switch. A no-holes modification can then be added to provide real cw selectivity at the expense of only a few components.

ham radio

the homebrew art

Bill Hayward, WØPEM, 3408 Monterey Street, St. Joseph, Missouri 64507

"The transmitter here is homebrew; the receiver here is homebrew." Remember when you used to hear that on the air? But today all you hear is, "The rig here is a Super Sky Model 10,000 Transceiver." Everyone then knows what it is.

What has happened to homebrewing?

First let's look and see what has happened to ham radio in the last 15 years. Single sideband, transistors, integrated circuits and the switch to transceivers have come into widespread use. But are these reasons not to homebrew? I don't think so. Now let's look at the marks against homebrewing: Takes time; can't find the parts; costs too much; doesn't look good; can't build ssb gear because I don't understand it; don't know anything about transistors. Let's take these one at a time and see how hard they really are.

time

It takes time; truly it does take time to do most anything, but if a few hours were set aside each week for a homebrew project, you could soon have it built. The more time you allow for the project the quicker it will get done. A receiver, transmitter or transceiver will take a lot more time than an swr bridge. So remember this when you start a project, and don't expect that receiver you might be building to be working in a day or two.

parts

This is sort of the same old thing, law of supply and demand. If you don't buy parts, the stores won't stock them. "Well then," you ask, "where can I get parts?" Allied Radio or Newark in Chicago, or Lafayette in New York, seem to have the best selection of parts. Sometimes Allied's regular catalog doesn't list some items, but their industrial catalog does. So if you build a lot it might pay to get an industrial catalog.

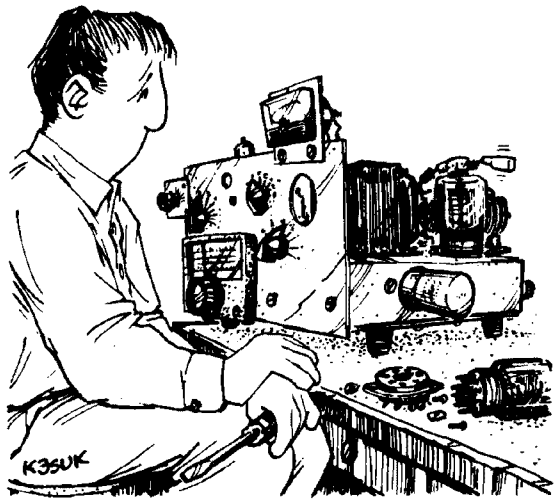
cost

The cost factor will vary, depending upon the size of your junk box. A homebrew rig doesn't have to cost a third of what a store-bought one would. If you don't have any parts to build with, maybe a trip to the local surplus store could help you. Sometimes a horsetrading trip to a local ham could net you some parts he's going to throw out. Either new or surplus parts will work, so the cost of homebrew equipment can be what you can afford.

appearance

A piece of homebrewed gear can look as good as you want it to. A little extra time and thought given to planning the layout, before any holes are cut, can make all the difference in the world as far as looks are concerned. Of course, the better a piece of equipment looks, the better it was laid out. So take that little extra time needed to make it look like a factory-built rig.

You can't build single sideband gear? If you've built a-m gear, you sure in the world can build ssb gear. Single sideband is much easier to build. Some will disagree



A little extra time and thought given to planning the layout before any holes are cut can make all the difference in the world.

with me here, but you don't have the big modulators, big power supplies, big cabinets, and big TVI problems with ssb that you do with a-m transmitters.

Single sideband transmitters use either the filter or phasing method, but the trend today is toward the filters.[1]

transistors

You don't know anything about transistors—you can't build with them? They're easier to wire than tubes because they have only three leads whereas a tube may have up to seven. The transistor requires no filaments so it operates cooler than a tube.

How can you find out about transistors? You might try some of the books on the market today. A good one is "Basic Theory and Application of Transistors;" it is TM-11-690 and costs only $1.25 from the Government Printing Office in Washington, D.C. It will give you a good basic understanding of transistors. Or you might try one of the new self-instruction programmed books on transistors. Maybe you could even take a course from one of the electronics schools in transistor theory. The ham who says he doesn't know about transistors doesn't **want** to learn about them.

design

You ask, "Where do I come up with a design?" I do R and C—Read and Copy. I start first with a list of things I want in a transmitter, then start looking through old magazines to see if I can find something close to what I want. Maybe I find one thing in one magazine; something else in another. I then combine the two into what I want.

Now that you have down on paper what you're going to build, you start looking through the junk box for parts, or maybe even make a trip to a friend's junk box. You find out then you don't have to order as many parts as you thought you would.

heat up the soldering iron

Then comes the big day when you start drilling holes. Make sure your layout is correct, because you are only building one transmitter, not four or five, and you can't afford to make mistakes.

Now that the holes are all drilled, you start mounting parts; then comes the wiring. Take your time here, because it'll pay off in the long run. If you get tired after wiring for a while, set the work aside for a day or two, then go back to it. But be sure and **keep notes.** They will come in handy when you go back to wiring.



Usually I find one thing in one magazine, something else in another.

the final check

It will take time and patience to get the equipment in proper working order. But the extra time spent on tuning up and alignment will make for a better-sounding transmitter. The final-final test, of course, will be the on-the-air reports.

The next time you're on the air, and someone wants to know what that good-sounding rig is, you can say with pride, "the rig here is homebrewed."

If you're thinking about buying a new receiver, transmitter, or transceiver, why not think of the homebrew route? Not only will you be saving money, but you can learn what goes on behind that panel. And, after all, isn't that what amateur radio really is?

reference

1. Fred S. Randall, K1UKX, "Homebrew 5-Band SSB Exciter," *ham radio*, March, 1968.

ham radio

propagation predictions for june

Victor R. Frank, WB6KAP, 12450 Skyline Boulevard, Woodside, California 94062

Perhaps you've judged from some of my past columns that I believe that there's much more to amateur radio than just communicating. This month's column is written for those amateurs who have a spark of the oldtime spirit of exploration of the unknown that brought amateur radio to its position of eminence. Let us revel a bit in these past glories.

The history of radio is replete with examples of contributions by amateur experimenters to science. In the beginning, there was Guglielmo Marconi. Although his later experiments were financed, and he was never a licensed amateur, his experiments were truly in the spirit and tradition of amateur radio.*

Marconi's transatlantic tests took place on wavelengths of 1000 to 2000 meters in 1901, in spite of the belief by most eminent scientists of the day that wireless communication was limited to the horizon. Kennelly and Heaviside independently postulated the existence of electrically conducting reflecting layers in 1902 to explain Marconi's results, and Balfour Stewart had predicted their existence more than 20 years previously to explain variations in the earth's magnetic field. However, the existence of these layers was not verified by experiments until 1925.

In the meantime many physicists were attempting to explain long distance propagation as being due to diffraction so they felt that the longer waves were the most useful. The Radio Act of 1912 restricted radio amateur operation to those "useless" wavelengths below 200 meters.

One of the earliest cooperative radio amateur scientific studies was the ARRL Bureau of Standards fading tests held on 250 meters in 1920. In 1921 and 1922 the ARRL sponsored successful transatlantic tests on 200 meters and in 1923, two-way transatlantic communications took place with Leon Deloy, 8AB, in France on 100 meters. The race to the short waves was on, and the science of ionospheric physics was born.

In the 1930's, correlations of the effects of meteorological conditions on vhf were made by Marconi in the Mediterranean and later by Ross Hull of the ARRL using radio amateur observations. Also, amateur reports of long-distance 5-meter contacts led J. A. Pierce, W1JFO, of Harvard University, to establish the cause as sporadic-E. Pierce also correctly identified meteors as being the cause of transcient hf propagation bursts inside the skip zone.

In the late 1940's, **CQ** magazine sponsored Radio Amateur Scientific Observations to map the motion of sporadic-E clouds using amateur 50 MHz observations.[2,3,4,5] Radio amateurs also contributed to science by reports of vhf reflections from the aurora.[6,7] Dr. O. G. Villard, Jr., W6QYT, at Stanford University introduced the radio amateur to such postwar developments as backscatter sounding,[8] meteor scatter communication,[9] artificial sporadic-E layers[10] and widespread use of single sideband.

Perhaps one of the greatest postwar amateur discoveries was that of trans-

* If you haven't read George Jacob's articles on Marconi[1] or the other excellent historical articles on "Men In Radio" in CQ magazine, you should. Also highly recommended are ARRL's historical articles, "Fifty Years of ARRL," which appeared in *QST* in 1964 and Clinton B. DeSoto's "200 Meters and Down."

equatorial propagation of vhf signals at night during years of high solar activity.[11] The National Bureau of Standards requested amateur observations of their South American 50 MHz beacon transmissions during the International Geophysical Year (IGY),[12] and the ARRL-IGY-PRP (International Geophysical Year Propagation Research Project)[13,14,15] brought together amateur observations of anomalous vhf ionospheric propagation from all over the world.

In 1957, J. Chambers, W6NLZ, and R. Thomas, KH6UK, demonstrated that 144-MHz tropospheric vhf contacts could be made over water as far as 2540 miles and in succeeding years repeated the feat on 220 MHz and 432 MHz (one-way). In 1959 and 1960, R. Mellen, W1IJD, and C. Milner, W1FVY, demonstrated that 50-MHz signals could be propagated for distances as great as 2500 miles from Ice Island T3 through the auroral zone to the northern contiguous United States.[16,17]

More recently, other than the OSCAR tests, amateur contributions to the state-of-the-art have left something to be desired. You may argue that, "These days true contributions to the state-of-the-art are made by large groups of well-financed researchers in government and industry; these days radio amateurs are at least ten years behind and many tens of dB's below the state of the art."

Perhaps you stand too much in awe of big technology. Sure, interesting discoveries **are** being made with facilities well beyond the capabilities of amateur radio. However, in over ten years of radio-propagation research I have been surprised by how much can be done by the properly motivated and educated "little guy."

Some of us just take up space on the bands; some of us **do** communicate; some of us do perform public service. A few of us may, once or twice in our lives, make a small contribution to science, or even an MSB (Major Scientific Breakthrough). At times most of us observe various geophysical phenomena—through frosted-over glasses. All that stands between just observing and actually contributing to science are perception, imagination, perseverance, and some skill, tempered with knowledge. For example, thousands of amateurs probably observed noise from outer space before Carl Jansky and Grote Reber, (W9GFZ),[18,19,20] but Jansky and Reber had the necessary combination of mental ingredients and technology to make the discovery of cosmic noise in the 1920's, and eventually to carry the observations through to start the science of radio astronomy.

Lately, I had the occasion to talk with a scientist of note who shall remain unnamed, and I asked him about using amateur observations of some upcoming ionospheric chemical releases. His reaction was negative, not for security reasons, but because he felt that the useful amount of information gained would be more than balanced out by the job of sorting through misinformation. One major factor for this was the lack of satisfactory prelaunch communications to amateurs on previous tests of this sort.

Would you believe it? If not, listen in to some W1AW broadcasts or Project OSCAR broadcasts. Soon you too will come to the conclusion that we are a long way from having a reliable 24-hour-a-day amateur radio communications system. And that is, unfortunately, the nature of the radio amateur service.

long delay echoes

There are still some scientists that have not lost faith in radio amateur observations. Dr. O. G. Villard, Jr., W6QYT, has recently enlisted the aid of the Northern and Southern California DX clubs in collecting reports of long delay echoes which may have been observed by radio amateurs in the hf bands. He is particularly interested in signals that last for two seconds or more, and is not interested in the frequently encountered round-the-world echo.

In 1928, C. Störmer in Norway heard echos with delays of 3 to 5 seconds on transmissions from a station at Eindhoven,

Netherlands on 30 meters operated by fellow ionospheric scientist B. Van der Pol. He telegraphed Van der Pol, who listened and heard the echoes himself. Störmer thought the echoes were due to reflections from solar ionized particle streams, while Van der Pol thought them due to great group retardation in the ionosphere.[21,22] Other scientists had other theories; E. V. Appleton thought the echoes may have originated from the signal circulating the earth repeatedly,[23] P. O. Pederson thought they may have been due to trapping of electrons along geomagnetic field lines,[24] and N. Janco thought they may have been due to trapping between ionospheric E and F layers.[25]

During the 1929 solar eclipse A. E. Kennelly observed echoes with delays between 1 and 30 seconds on 25 meters.[26] Since 1934, reports of long delay echoes have been infrequent and have come only from radio amateurs. One systematic attempt to study the echoes was made in 1947 at Cambridge University in England, but it was unsuccessful.[27] For this reason, and the unsatisfactory possible explanations of this effect, most scientists today are inclined to discount the observations as imaginary.

However, the situation has changed within the past two years. Laboratory experiments with plasmas by Dr. F. W. Crawford at Stanford have indicated that microwave signals may be stored in a plasma and retrieved after an appreciable time delay. If the laboratory results could be extrapolated to the ionosphere, the delays would be tens of seconds and thus there is an awakened interest in long delay echoes.

Dr. Villard would like to hear from radio amateurs who have heard echoes of this sort on their or other station's signals with as much of the particulars as can be remembered. Tape recordings of such an effect are most welcome. The desired information includes call signs, addresses, date and time, frequency, modulation type, transmitter power output, antenna type, gain, direction of fire and height above ground, and estimated duration of echo and characteristics of same, such as Doppler shifts, fading, strength or multiple echoes. In addition Dr. Villard would be interested in knowing your occupation and the year you obtained your first license. Reports should be sent to:

Dr. O. G. Villard, Jr.
Radioscience Laboratory
Stanford University
Stanford, California 94305

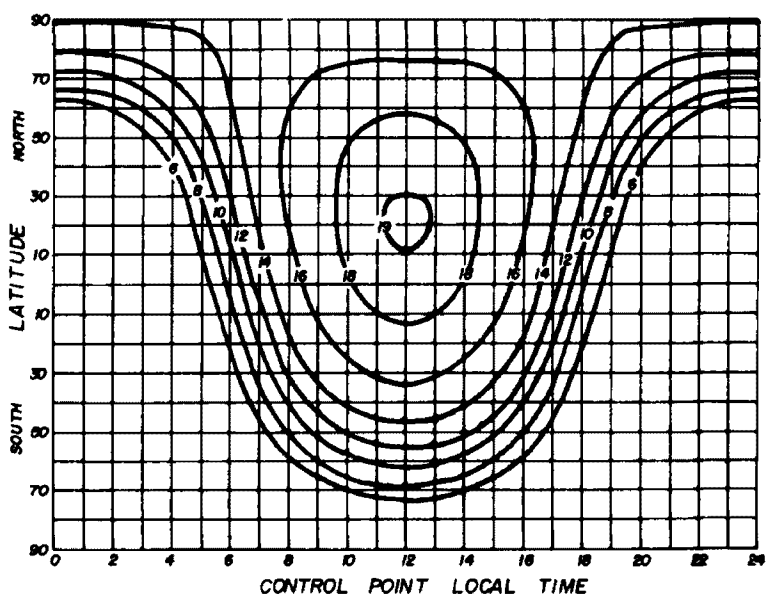
sporadic-E

I am interested in obtaining reports of 144-MHz sporadic-E and reports of 50-MHz sporadic-E under 400 miles at times when 144 MHz was open for sporadic-E last summer. I am also interested in reports of 50-MHz multiple-hop sporadic-E over distances greater than 2800 miles.

six-meter propagation during february and march

On February 2nd from 2215 to 2300Z, six meters was reported open for F2-layer backscatter between Northern and Southern California, Arizona and Kansas. WB6NTL in Sacramento reportedly worked a KP4 near 2230Z. On February 14th, ZB2BC, Gibraltar, reportedly heard ZE1AZG (beacon

fig. 1. Time chart of median E-layer for June 1969 predicted for the regular daytime E-layer.



station at Port Victoria, Rhodesia on 50.050) near local noon.

Solar activity increased drastically during the last week of the month. From February 20 to 27 the daily sunspot number was in excess of 200 and as high as 300, up substantially from earlier weeks. An active region on the sun near 12 degrees N. Latitude produced several solar flares of importance on February 24th at 2300Z, on February 25th at 0900Z, on February 26th at 0421Z, and on February 27th at 1358Z. Severe ionospheric disturbances followed the flares of the 25th and 27th.

Ionospheric physicists have found that the particles emitted from the sun during a solar flare are confined to streams and that the probability of the earth intercepting the stream from any given flare is small. Once a stream has bridged the gap from sun to earth, however, it forms a well conducting track through which subsequent flares may more easily shoot particles. This appears to have been the case during the period of February 27 through March 2, 1969.

On February 27th, W5RCT near Dallas, Texas reports working XE1AAN and hearing ZK1AA's beacon on 50.100 between 0135 and 0315Z, peaking as high as 40 dB over the noise at 0200Z and fading down into the noise between 0232 and 0259Z. March 1st was the beginning of the transequatorial season for CE3QG at Santiago, Chile when he worked into Colorado, Texas, New Mexico and Arizona. Sporadic-E was also present between these states and California. In addition, some fuzzy-sounding signals from locals were heard out of the northeast that resembled aurora. Most of this was between 0400 and 0600Z on March 2.

On the afternoon of March 8, there was a 50-MHz F2 opening between California and Hawaii for three hours with KH6EQF reported, "very loud." Backscatter was also reported between Northern and Southern California.

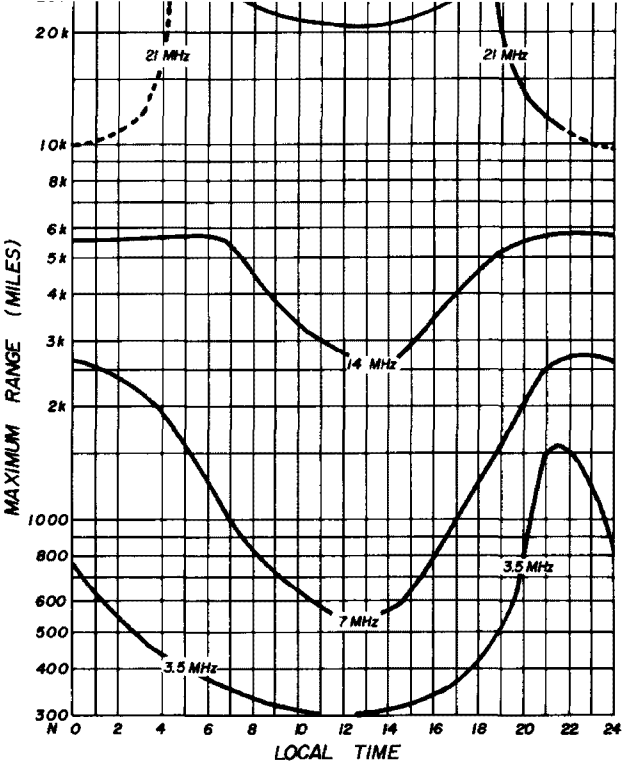
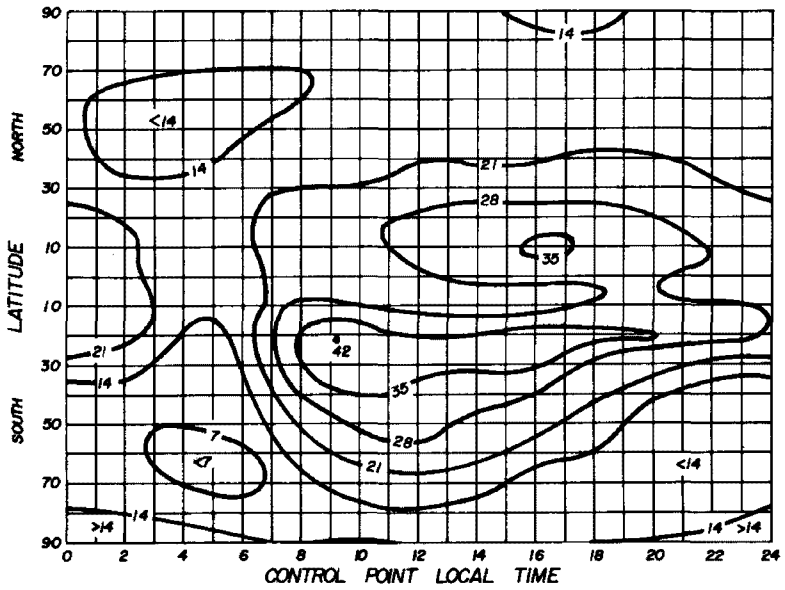


fig. 3. Maximum range to the north from 38° N. latitude as limited by absorption, atmospheric noise and system parameters (100 watts CW to typical antennas).

fig. 2. Time chart of median F2 muf for June 1969 from ITS predictions centered on a longitude of 90° W.



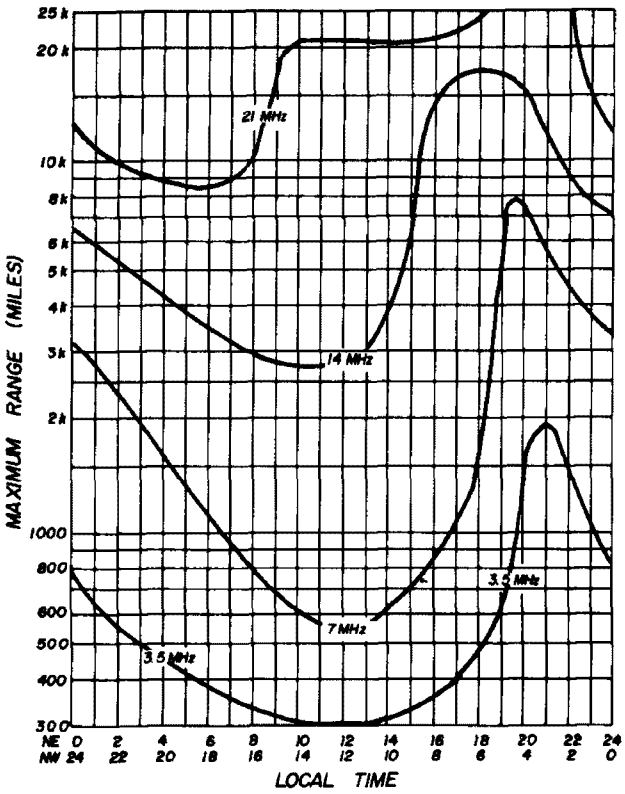


fig. 4. Maximum range to the northeast (top time scale) and to the northwest (bottom time scale) from 38° N. latitude.

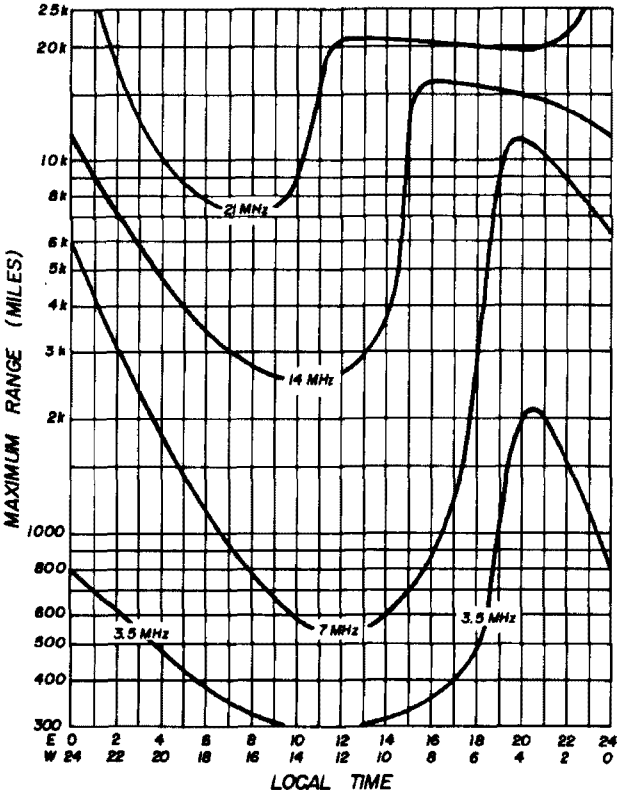


fig. 5. Maximum range to the east (top time scale) and to the west (bottom time scale) from 38° N. latitude.

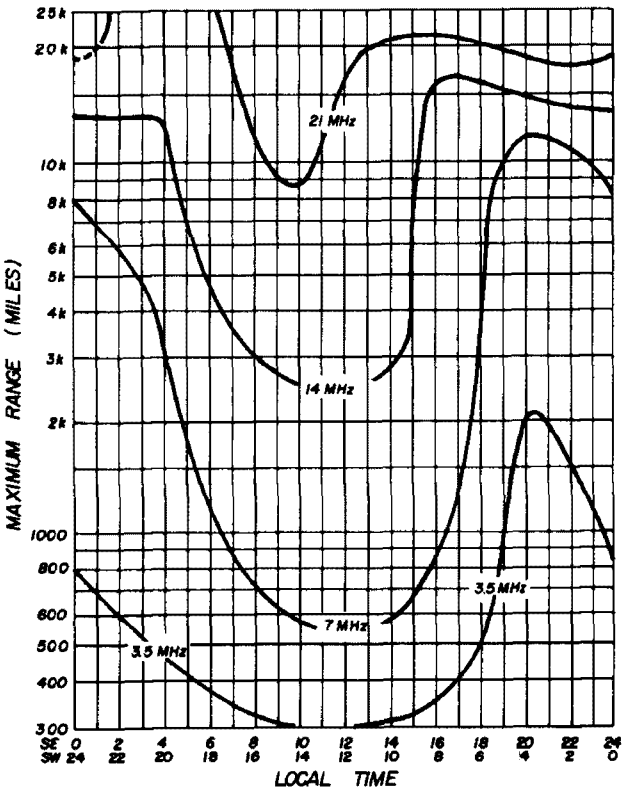


fig. 6. Maximum range to the southeast (top time scale) and to the southwest (bottom time scale) from 38° N. latitude.

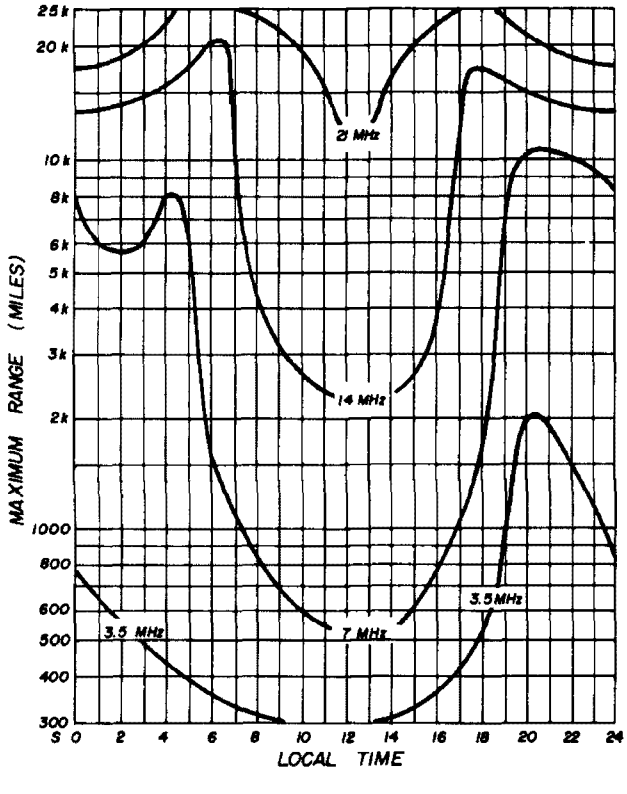


fig. 7. Maximum range to the south from 38° N. latitude.

how to use these propagation charts

1. To find the maximum usable frequency for F2-layer propagation for distances of 2500 miles or more in any direction, find your control point and read the frequency from the F2-layer muf time chart. Your F2-layer control point is 1200 mies away from your station in the direction of propagation; this is about an 18-degree difference in latitude for a north-south path, or 1½ hours time difference for an east-west path.

2. To determine the path muf for a path under 2500 miles, the control point is at the path midpoint, and a correction factor is applied to the 2500 mile muf. These correction factors are plotted in fig. 7 of the August 1968 column.

3. During summer daytime, the path muf for a path shorter than 1200 miles may be set by E-layer propagation. To determine if this is so, refer to the E-layer muf time chart and the chart of muf reduction factor vs distance in this month's column. Note, however, that sporadic-E will probably result in zero skip distance on 7 MHz and skip distances under 400 miles on 14 MHz very frequently during daylight and evening hours.

4. The F2-layer will probably be effectively shielded by the E-layer for operating frequencies below 70 percent of the predicted E-layer muf.

5. Over any particular path involving more than one hop, the path muf is the lower of yours and the other station's control-point muf. The muf time charts may be treated as muf contour maps. (The F2-layer chart has significant errors outside the range of longitudes between 45°W. and 135°W.) As such, each hour is the equivalent of 15° of longitude. A map drawn to the same scale can be overlayed and positioned to the right or left to show the variation of the muf contour map with time. Curved lines may be drawn on the overlay representing great circle paths, as found from a globe or "Ionospheric Radio Propagation," printed by the U. S. Government Printing Office.

6. To find the maximum propagation distance as limited by ionospheric absorption and atmospheric noise, refer to the maximum range charts for the directions *you wish to* work. Note that the time scales are reversed for westward propagation. Also note that this month the noise curves have not been assumed to be symmetrical about local noon. Thus when the curves are used to predict propagation to the west they may be in error due to this lack of symmetry. These curves are based on a unity signal-to-noise ratio of a 6-kHz bandwidth with 100 watts cutput and antenna gains (over an isotrope) of 6 dB for 20 and 15 meters, 0 dB for 40 meters, and —6 dB for 80 meters at each station.

propagation for june

Summertime conditions described in last month's column will be with us for several more months. In short, this means more noise on 80 and 40 meters and 20 meters at night, and more absorption on the same bands during the daytime. The inability to do much DX'ing on these bands near midday will be due as much to shielding of the F2-layer by the E-layer as due to absorption.

It takes a lot of E-layer hops to cover more than transcontinental distances at respectful elevation angles, and each hop gets absorbed more than a longer F2-layer hop. As a result best DX will be during twilight and darkness hours.

Although 15 meters is primarily a daytime band it may remain open well towards local midnight. Some mornings the MUF may not quite reach 21 MHz, or all you may be able to work is South Americans. You should be able to work polar paths that do not reach latitudes higher than 60 degrees north latitude for a fair fraction of the days of the month. Don't look for much F2-layer propagation between the East and West Coasts on 28 MHz during the month. There will be double-hop and sporadic-E occasionally, and regular openings to the Southern Hemisphere, however.

June is the big month for sporadic-E on 50 MHz (and perhaps on 144 MHz). In addition,

there are daylight meteor showers scheduled for June 4th-6th (Perdeids), June 8th (Arietids) and June 30th-July 2nd (Taurids), and the nighttime showers, Scorpids (June 2-17) and Pons Winnecke (June 27-30). Judging by the connection shown between sporadic-E and meteor ionization in last month's column, these dates might bear watching for widespread sporadic-E as well.

The F2-layer muf time chart was derived from "Ionospheric Predictions" for a longitude of 90° W. These predictions are published monthly by the Institute of Telecommunications Sciences (ITS), Boulder, Colorado, and are available through the U. S. Government Printing Office. The maximum distance curves were derived from consideration of atmospheric noise levels (from CCIR Report 322) and calculated path losses at fixed distances in each direction from 38° N. latitude. Some minor differences in maximum range will be noted due to change in absorption and change in noise levels throughout the 48 contiguous United States. Somewhat greater ranges should be expected over paths further from the sub-solar point (more northerly latitudes during daylight).

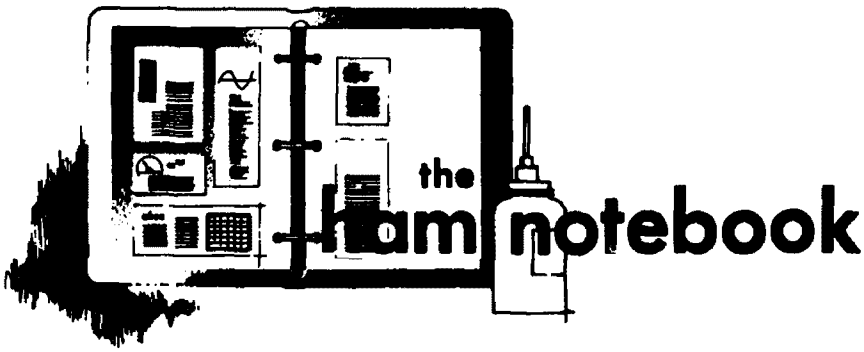
The predictions given in this column are for median conditions. On any particular day, muf's may be as much as 10 percent higher or lower than the median. Absorption and noise levels, particularly on the lower amateur bands, may be as much as 10 dB different from the median. Residential noise levels (from electrical lines, appliances, and vehicular traffic) may, and frequently will, be tens of decibels stronger than atmospheric noise.

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Many of the references concerning radio amateur work in radio science prior to the Second World War have not been cited because they are not generally available and because the historical summaries in more recent issues of *QST* and *CQ* adequately cover this early period.

ham radio



tone encoder and secondary frequency oscillator

While working on various RTTY and fm selective-call projects, the need for an accurate stabile tone source became apparent. The device shown here is presently programmed for eight frequencies, provides a good sine-wave output and is quite stable over a wide range of temperature, voltage and loading. The oscillator has been duplicated by a number of amateurs and used as a digital selective-call encoder on many of the West Coast repeaters. Its reliability and stability have been proven under all types of field conditions. Further, the unit can be easily modified to provide tone-burst operation.

It should be obvious that to obtain frequency accuracy closer than 1%, a frequency counter must be used. However, close to 1% accuracy can be expected for the frequencies listed with combinations of good quality Mylar capacitors.

For burst use, return the negative common to the push-to-talk line instead of ground; add a zener regulator for power or use a dry battery. Connect the two diodes and the RC network as shown in the box in **fig. 1.** For mobile operation, be sure **not** to ground or

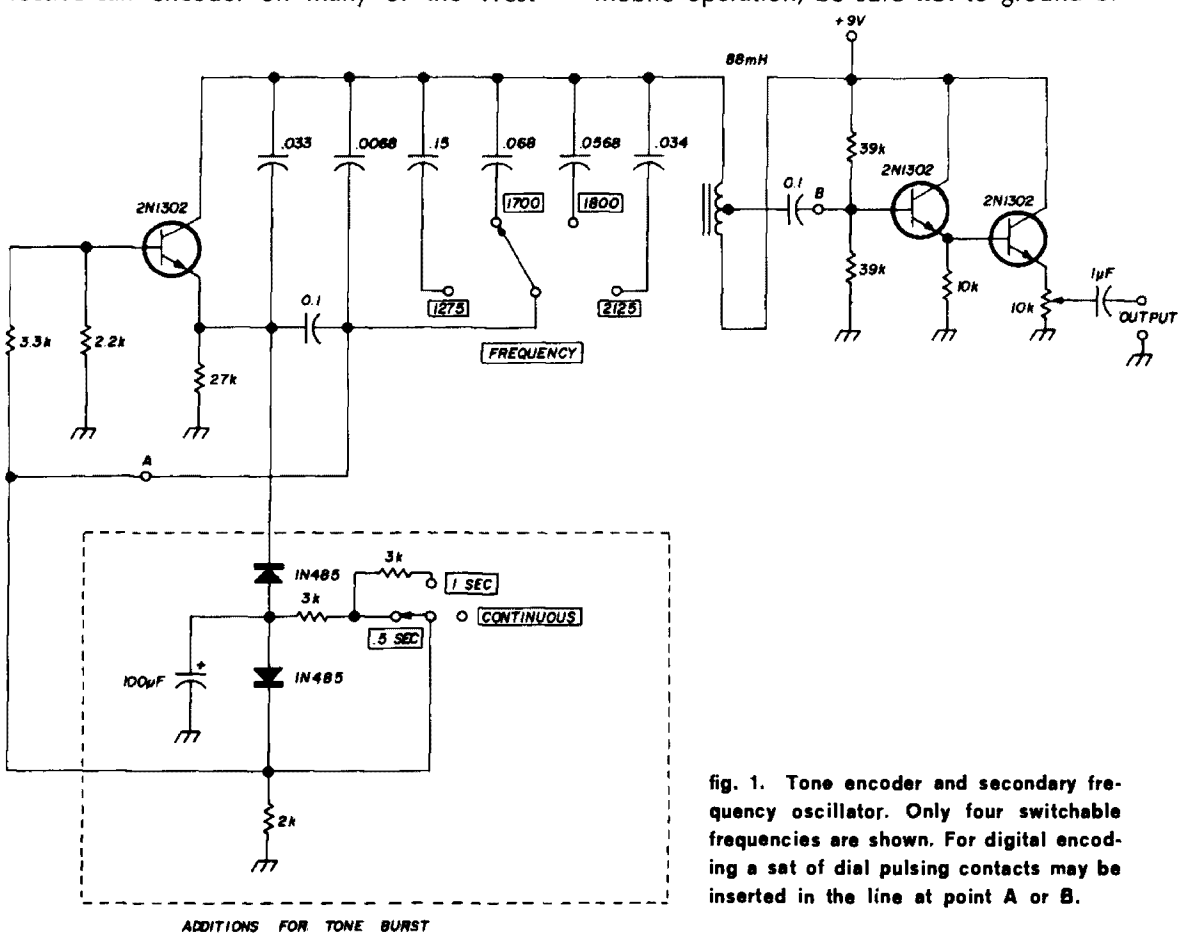


fig. 1. Tone encoder and secondary frequency oscillator. Only four switchable frequencies are shown. For digital encoding a set of dial pulsing contacts may be inserted in the line at point A or B.

table 1. Frequencies and tuning capacitors used in the tone encoder built by WB6YQC.

Channel	Design Frequency (Hz)	Measured Frequency (Hz)	Capacitance (μF)	Use
1	1275	1271.5	0.150	RTTY space frequency*
2	1700	1699.6	0.068	RTTY mid-band*
3	1800	1803.0	0.0568	selective-call tone
4	2125	2125.8	0.0340	RTTY mark frequency*
5	2400	2401.4	0.0181	selective-call tone
6	2550	2551.4	0.0106	RTTY mid-band
7	2700	2704.1	0.0071	selective-call tone
8	2975	2973.3	0.0001	RTTY space frequency

*wide shift fsk/ssb transmitter.

common **anything** except through the audio return line to the transceiver.

Digital encoding can be easily obtained by inserting a set of dial pulsing contacts in series with the 3.3k oscillator base-bias resister, or in series with the 0.1 μF oscillator output coupling capacitor. The unloaded output voltage is in excess of 2 V rms above 1700 Hz. With a 100-ohm load, the output drops to 300 mV rms, still enough to provide full deviation in most fm rigs.

The frequencies I used are shown in **table 1** but other tones can be easily substituted for your particular requirements. Printed circuit boards for the basic oscillator circuit are available from the treasurer of the Grizzly Peak VHF Radio Club, Post Office Box 1333, Richmond, California 94802 for $1.50 each. I would like to extend my thanks to WA6UFW who did the initial design work with the basic oscillator.

David M. Stahley, K8AUH

grounding

If you're plagued by excessive receiver noise, a hot microphone, or rf around the shack, better check your ground connection. If your house was built after about 1960, you may be lacking a simple metal-to-metal connection if you depend upon a cold water pipe for a ground. Most builders have been using various forms of polyvinyl pipe to bring water in to a given location in the house where it is connected to standard metal fittings. The contractor saves a pile of money with this approach but you haven't got a ground; play it safe and use a good ground rod.

Elliott Kanter, W9KXJ

incremental tuning for the heath HW-100

The addition of incremental tuning to a ssb transceiver is an ideal and often needed refinement. It allows the receiver to be shifted a few kHz above or below the transmitting frequency. This helps when someone has trouble zero-beating your frequency or when you're in a roundtable where someone is inevitably off frequency. It can eliminate the embarrassing problem of tuning for the off-frequency station and then suddenly finding yourself equally off zero beat. It is especially handy where there may be a highpitched voice—you can tune for personal listening comfort without affecting your own transmit frequency.

In the receive position, with the incremental tuning switch on, varying the dc bias on the varicap diode causes the capacitance to change within the range of the incremental tuning control. With the transceiver in the tune position (or transmitting), the antenna changeover relay (through contacts 3, 7 and 11) disconnects the manual tuning voltage and inserts a calibrate voltage in its place. This sets the capacitance of the varicap at the center frequency position of the incremental tuning control. A shift of 2 kHz either side of center can be obtained with the values shown in the schematic.

The bias voltage is derived from the +150 V regulated supply. The voltage is dropped to +12 V and regulated by the 47k, 1-watt resistor and a 1N963B zener diode; this keeps the voltage steady on both transmit and receive. The photo shows how the components may be mounted in an already well packed area.

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The calibrate pot is mounted just to the side of the final tank cage in an existing 5/16-inch hole. (This hole will have to be enlarged to 3/8-inch to accommodate most pots.) Mount a 2-lug terminal strip (Cinch-Jones 52) on the underside of the chassis near the pot. A 47k, 1-watt resistor is placed between pin 1 of the OA2 (V18) to the closest tie point with the zener diode going from the same tie point to ground. The incremental tuning pot on the front panel is mounted to the right of the meter at the same height as the tuning controls to maintain a neat panel layout. You can get an extra knob from Heath to match the original equipment (part number 462-191 $.70). A sub-miniature spdt switch is mounted above the pot. Decal numbers may be added to indicate 1- and 2-kHz deviations from the center.

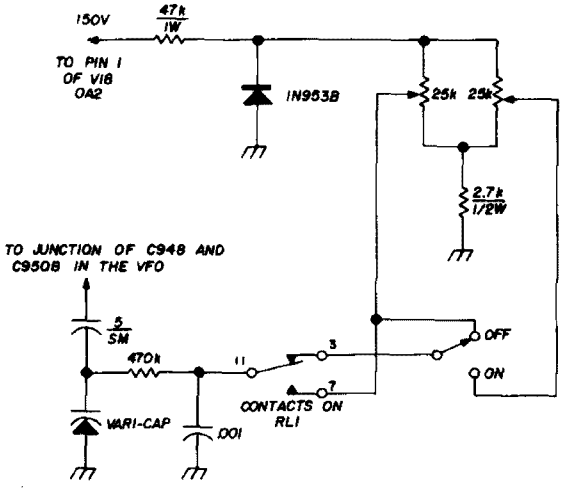


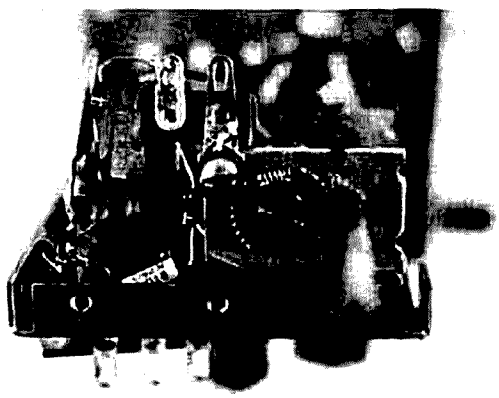
fig. 2. Receiver incremental tuning for the HW-100. The two 25k pots should have linear tapers.

A five-conductor color coded cable can be made up to wire between the chassis and the front panel controls. This cable can be passed through an existing 1/4-inch hole in the i-f circuit board.

Mount a one-terminal Cinch-Jones 51A terminal strip to the left of the backlash gear on the vfo capacitor; straighten the ground lug. Mount the varicap from the ungrounded lug to ground and mount the 5-pF mica capacitor from the same un-

grounded lug to the junction of C948 and C950B.

Mount a two-terminal Cinch-Jones 52A terminal strip just in back of the vfo capacitor on the stud holding the vfo coil. Mount the 470k, ½-watt resistor between ungrounded terminals, by-passing one end of the resistor with a .001 disc ceramic to ground and connecting a wire between the other end of the resistor and the terminal strip mounted on the vfo. Leave enough slack in the rigid wire to clear the rotor. Connect a wire to the bypassed end of the 470k resistor and pass it through a hole in the top of the vfo chassis to a Cinch-Jones 53 terminal strip mounted on top.



Varicap installation.

To calibrate the two pots for the same receive and transmit frequency, set the front panel pot (number 1) to its center position. Set the number 2 pot (mounted near the final tank cage) to its approximate center. With the incremental tuning switch turned on, set the mode switch to CW and tune the vfo dial to zero beat the crystal calibrator. Put the incremental tuning switch in the off position and adjust the number-2 pot to zero beat without touching the vfo dial. With this calibration completed you should receive and transmit on the same frequency. Some minor correction in dial accuracy will have to be made to allow for slight added capacitance, but it is minor.

Bill McCracken, K1GUU

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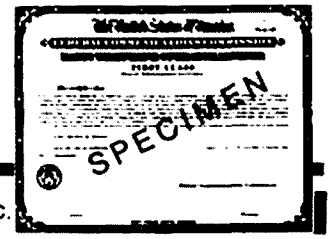
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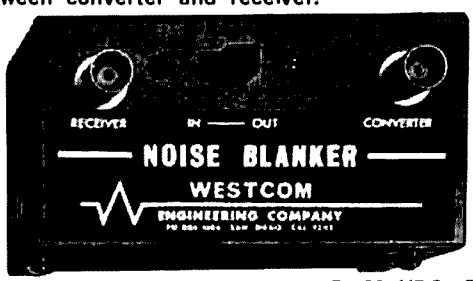
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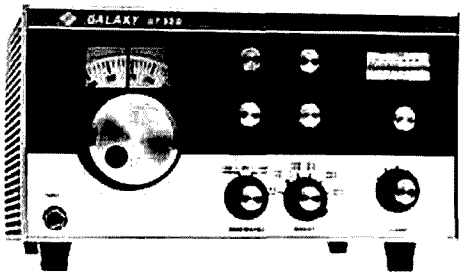
new products

two-meter fm transceiver

Two-meter fm is becoming very popular in many areas, and new two-meter fm repeaters are popping up all over the country. A new entry in the fm transceiver market is the FDFM-2S manufactured by the Inoue Communications Equipment Corporation. The FDFM-2S is completely solid state and built around modular computer techniques so it's rugged and has good interstage shielding. The unit features three sets of transmit/receive frequencies that are selectable from the front panel; power supply is 12 to 24.5 Vdc so it's a natural for mobile operation.

The receiver is dual conversion with a fet rf amplifier and adjustable squelch; sensitivity is less than 1 μV for 20 dB signal-plus-noise-to-noise ratio, and spurious responses are 60 dB down. The transmitter section runs 10 watts input (5 watts out), has frequency deviation less than 25 kHz and built-in push-to-talk. The FDFM-2S comes complete with microphone, battery pack, six crystals of your choice, mobile mounting bracket, power connectors and manual; $295.95. Also available are the 2-watt FDFM-2 two-meter fm transceiver and 2-watt FDAM-3 six-meter a-m/fm transceiver with vfo. For more information write to Varitronics Incorporated, 4109 North 39th Street, Phoenix, Arizona 85018.

new galaxy transceiver



Galaxy Electronics has announced a completely new transceiver, the GT-550, and a complete line of matching accessories. The GT-550 is a five-band ssb and CW transceiver that features 550 watts input on ssb (360 watts input on CW). Optional accessories include an ac power supply, mobile power supply, phone patch, cw filter, vox, crystal calibrator and mobile mounting bracket. The new Galaxy line of matching equipment includes the LA linear amplifier, rf console, remote vfo and speaker console. Price of the basic GT-550 transceiver is $449. For an illustrated brochure with full specifications on the complete Galaxy line, write to Galaxy Electronics, 10 South 34th Street, Council Bluffs, Iowa 51501.

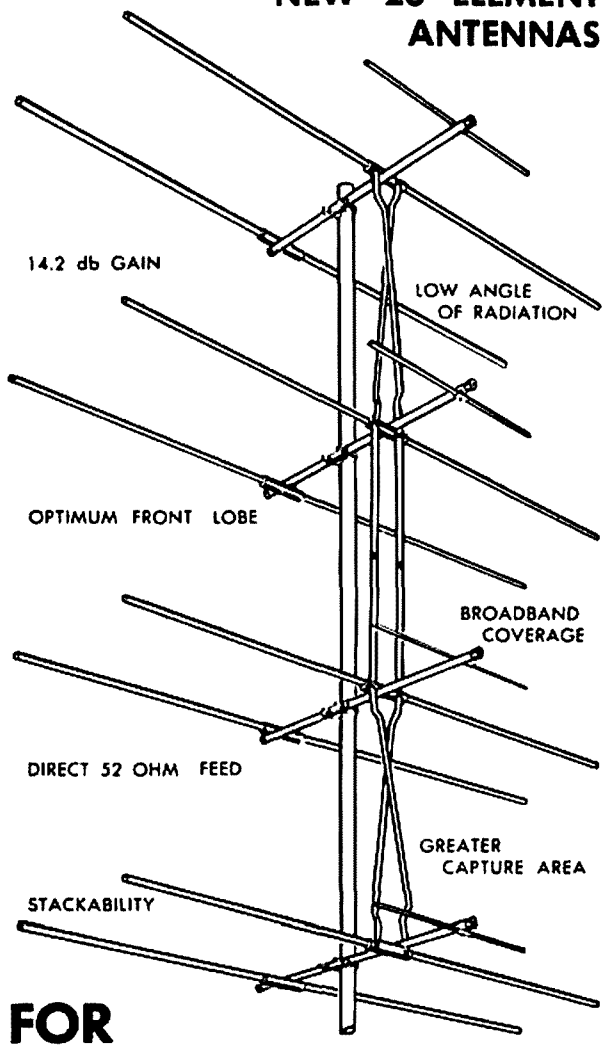
new arrl handbook really revised

If you're one of the thousands who found out about the 1968 ARRL Handbook too late (It was sold out by August; they usually expect to have copies left in February), you'll be happy to hear that the 1969 edition is now available. Last year's **Radio Amateur's Handbook** was extensively revised by Doug DeMaw, W1CER, to incorporate modern solid-state circuits and practices. Its success proves both the need for revision and its satisfaction. The new 1969 edition is even better. Apparently, as is necessary in such a comprehensive work, the theory is being revised a few chapters at a time, while practical construction projects are changed more easily.

The big news is that in addition to the excellent basic chapter on semiconductors, other chapters incorporate semiconductor

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circuits (along with vacuum-tube circuits). The receiving chapter includes bipolar, fet and IC detectors, mixers, oscillators, rf and i-f amplifiers, Q multipliers and other circuits. Specific construction projects are included. The transmitting chapter has not been revised as extensively—yet—but a number of practical transistor transmitters and vfo's will interest you. The audio section now includes many solid-state circuits, and the code chapter an IC keyer.

The chapter on power supplies incorporates the information required to design reliable high-voltage power supplies with silicon diodes, and a number of low-voltage supplies are included for transistor gear. The vhf receiving chapter is outstanding. It reflects the latest amateur practice: fet's, bipolar transistors, even IC's. There's not a project using vacuum tubes in it! The simple 6- and 2-meter converters are excellent successors to the nuvistor converters that were the mainstays of vhf hams for many years, and the preamp for 23 cm should satisfy the real experimenter.

The vhf-transmitting chapter includes a clean-looking practical varactor tripler for 70 cm, and, it should be noted, efficient vacuum tube amplifiers for 70 and 23 cm. (I don't know of any transistors good for 500 W at 432 MHz!)

The section on portable and mobile gear is full of simple solid-state gear that should interest any ham: transceivers for 6 and 2, and a transmitter and converter for 160 m. Other chapters also include semiconductor gear. This discussion of the solid-state content of the handbook is simply intended to emphasize the extensive revisions Doug has made to incorporate modern practice.

The book also covers much other new material, plus the invaluable data that has always made the ARRL handbook a necessity in every ham shack. Now, moreover, it appears that a new handbook every three or four years isn't enough. The 1969 edition is a new book; even if you have a '68 copy, you'll want it. $4 from the American Radio Relay League, 225 Main Street, Newington, Connecticut 06111, or try your local distributor.

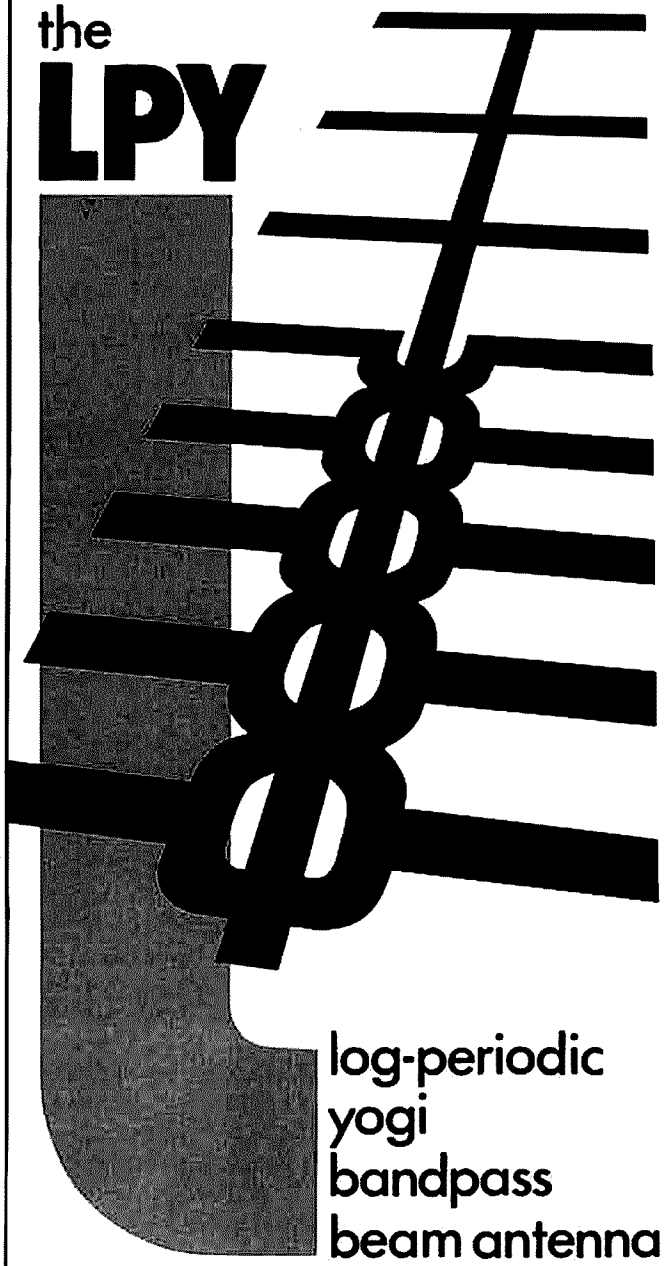
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JULY 1969

the LPY



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staff

editor
James R. Fisk, W1DTY

roving editor
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associate editors
A. Norman Into, Jr., W1CCZ
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art director
Jean Frey

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T. H. Tenney, Jr. W1NLB

offices

Greenville, New Hampshire 03048
Telephone: 603-878-1441

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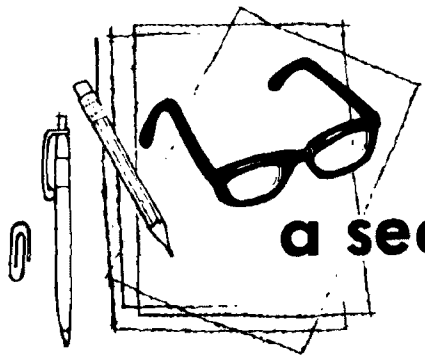
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a second look

by jim fisk

One of the more worthwhile amateur radio projects I have heard about recently is the John W. Gore Memorial Scholarship sponsored by the Foundation for Amateur Radio. The scholarship for 1969 consists of a $500 award for either graduate or under-graduate study.

Licensed radio amateurs who intend to make their career in electronics or related sciences may request a scholarship application. To be eligible, applicants must have completed one year of an accredited program at an accredited college or university leading to a bachelor's or higher degree. They must also hold a valid FCC license of at least General class. Although preference will be given to applications from the area served by the Foundation for Amateur Radio—the District of Columbia, Maryland and Virginia, those living elsewhere are not excluded.

Scholarship applications should be completed and mailed not later than August 15, 1969, and should be sent to the Chairman, Scholarship Commitee, Foundation for Amateur Radio, Inc., 449 Greenwich Parkway, N.W., Washington, D. C. 20007. The recipient of the scholarship award will be announced on September 1st.

The Foundation for Amateur Radio, Inc. is a non-profit organization devoted to the advancement of amateur radio and is composed of trustees representing over 20 radio clubs in the District of Columbia, Maryland and Virginia. John W. Gore, W3PRL, in whose honor the scholarship is named, was, until his death in 1960, the president of the Foundation.

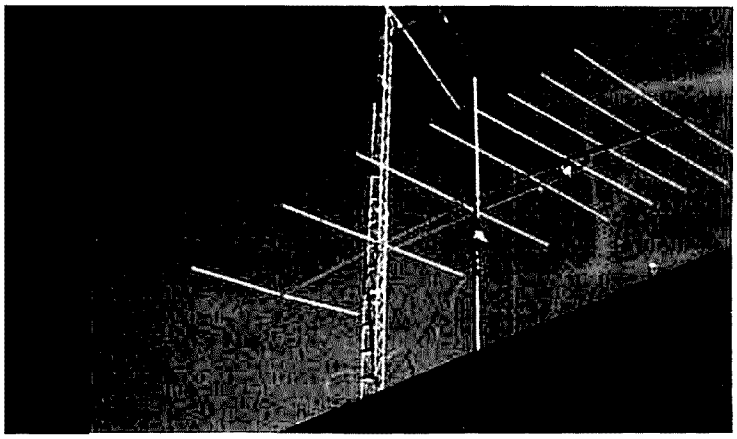
bonadio antenna

The article on the Bonadio antennas published in the April issue has received everything from brickbats to bouquets. One irate reader wrote in to say that, "Antennas and other equipment items are subject to measurements and reason; their operation must be explained and tested within the realm of physics—not metaphysics." More enthusiastic readers were unhappy when complete construction details didn't appear in the next issue—one amateur even had the support poles ordered for the space-dimension antenna.

Initial response to the Bonadio antennas has been for the most part emotional and the letters I have seen have contributed little to their understanding. Whether they work or not, I don't know, but W2WLR claims that they do. He has been trying for nearly 15 years to get something into print—now that it's in print, let's have some controlled experiments to determine if these antennas have any merit or whether the performance he is getting is a result of his particular location. Amateurs seriously interested in constructing one of these antennas can obtain more detailed constructional information from either W2WLR or myself.

An antenna that is apparently related to the Bonadio antennas was used with early English radar systems, type TRU. The TRU antenna consisted of two half-wave radiators, arranged in a cross configuration similar to the Bonadio square-diagonal arrangement, and fed through a Helmholtz coil. According to WØKWL they were able to track small bombers over 350 miles away without mechanically rotating the elements; frequency of operation was about 30 MHz. If any of you have more information on the TRU system—and particularly the unusual antenna—I would like to hear more details. The system is probably declassified by now, so perhaps someone even has an instruction manual they'd be willing to loan to me.

Jim Fisk, W1DTY
Editor



log-periodic yagi beam antenna

An adaptation of the Yagi array to obtain better bandwidth performance

Del Crowell, K6RIL, 1674 Morgan Street, Mountain View, California 94040
William Orr, W6SAI, Amateur Service Department, Eimac Division of Varian, San Carlos, California 94070

The Yagi parasitic beam antenna consists of a driven element plus a number of parasitic elements that increase the gain or directivity of the radiation pattern over that of a dipole antenna.[1] The number of parasitic elements, their tuning, and their spacing with respect to the driven element determine the characteristics of the array.

Generally speaking, the Yagi provides the greatest gain per unit size of any antenna array. Under normal circumstances, the more the elements, the greater the gain and the sharper the pattern of the Yagi. But as the number of elements increases, the more restricted will be the bandwidth of this popular antenna.

This bandwidth restriction, increasingly critical with respect to antenna gain, is of minor importance in the high-frequency bands, which are narrow, and where the Yagi is of modest size. However, antenna bandwidth becomes of paramount importance in the wider amateur bands where Yagi arrays are larger with respect to wavelength.

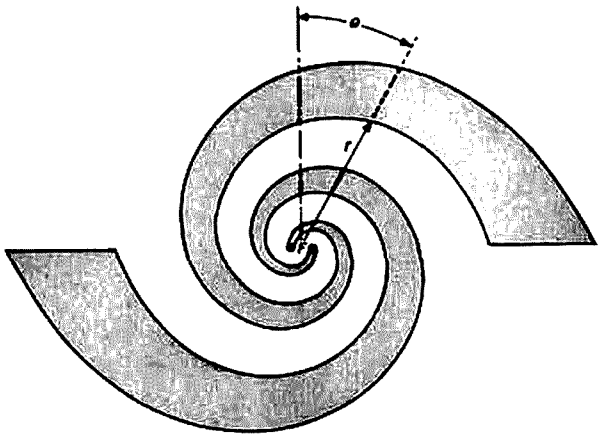
equiangular antenna concept

The restricted bandwidth of the Yagi beam may be improved by abandoning the parasitic element approach and applying instead equiangular principles. The equiangular principle deals with the design and assembly of frequency independent radiators. It is based on the unique idea that if the shape of an antenna can be specified entirely by angles, antenna performance would be independent of frequency.[2]

Practically speaking, this means that if all the dimensions of a radiator are

scaled by a constant factor, the physical size of the antenna may be changed without changing any of its electrical characteristics, provided its operating wavelength is changed by the same amount.

fig. 1. Equiangular spiral antenna is symmetrical about foedpoint F and is described in terms of angle θ and radius r from polar axis.



A simple two-dimensional, equiangular spiral antenna that conforms to this requirement is shown in **fig. 1.**

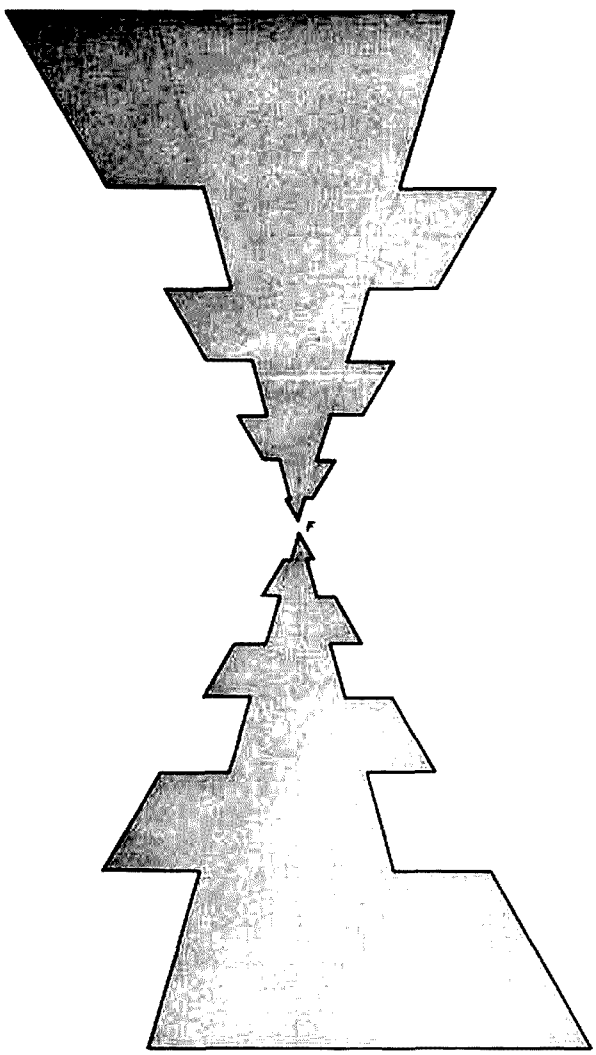
To be truly independent of frequency, the equiangular spiral antenna would have to start at an infinitely small point and expand to infinity. Practically, the antenna must have a feed point of finite dimension at the center and must have outer limits, as it cannot be infinite in size. Thus the frequency coverage of the structure is finite and is defined by physical, not electrical limitations.

Within the frequency limits imposed by these practical considerations, the equiangular spiral antenna resembles a frequency independent structure. This comes about because the transmission line is equivalent to the missing center portion of the structure, and the truncated outer portion does not affect the electrical properties to any significant degree. Most of the energy is radiated before it reaches the end of the structure if the antenna is large compared to the size of the radio wave.

the log-periodic antenna

The equiangular spiral antenna has limited use and exhibits little power gain over a dipole. A modification of the equiangular spiral antenna that provides power gain and directivity is shown in **fig. 2** wherein a planar structure is repeated periodically with respect to the logarithm of the frequency.[3] An antenna array of this configuration has characteristics that change with frequency, but before the change is very great in terms of wavelength, the structure **repeats itself.** A combination of the equiangular approach with the concept of periodicity results in a

fig. 2. Periodic planar antenna. Structures may be bent back upon themselves to form a three-dimensional array.



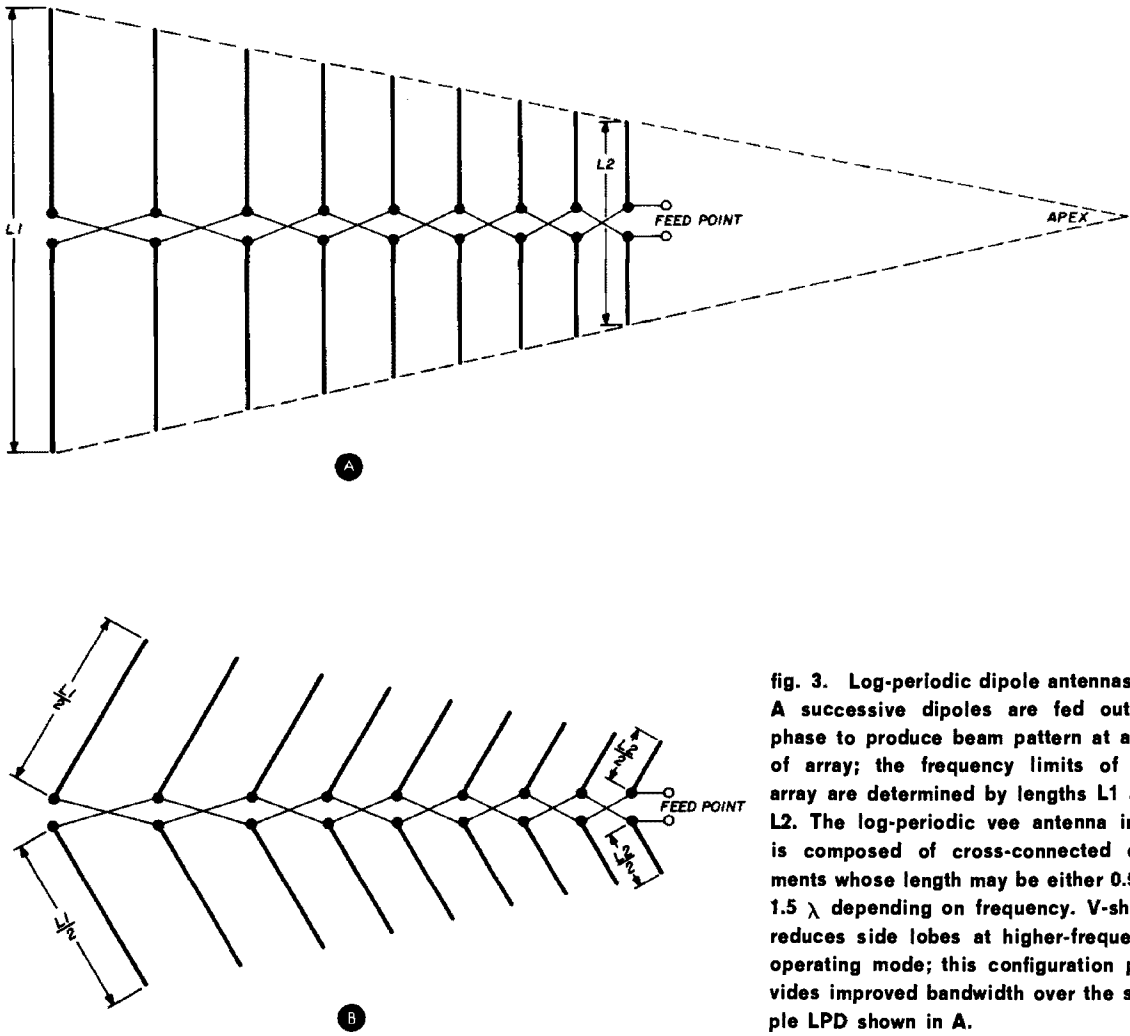


fig. 3. Log-periodic dipole antennas. In A successive dipoles are fed out of phase to produce beam pattern at apex of array; the frequency limits of the array are determined by lengths L1 and L2. The log-periodic vee antenna in B is composed of cross-connected elements whose length may be either 0.5 or 1.5 λ depending on frequency. V-shape reduces side lobes at higher-frequency operating mode; this configuration provides improved bandwidth over the simple LPD shown in A.

practical antenna with directivity and power gain and which has a bandwidth limited only by the antenna's physical size.

Various forms of the log-periodic antenna have been designed for specific uses. One of the most popular designs is the **log-periodic dipole** (LPD) array. The LPD antenna consists of a series of dipoles fed at the center and connected to the opposite wires of a balanced transmission line (**fig. 3**). The dipole lengths are consecutively shorter, and radio energy travels along the transmission line until it reaches a portion of the dipole structure where the length of the dipoles and their phase relationship combine to produce radiation.

The radiation is directed along the array toward the apex, so that the shorter elements either tend to act as directors or, if very short, are inactive. The LPD array can be fed at the apex with a balanced line, or with a balun and coaxial line. The antenna performs over a frequency span with limits defined by frequencies at which the extreme elements of the configuration are about one-half wavelength long. Antennas of this general type are commonly used for tv reception and are often used for amateur work at vhf.[4]

frequency limitations of the yagi

The Yagi parasitic beam functions as a directional antenna having power gain by virtue of the proper phasing of the parasitic elements. For high-frequency Yagi beams commonly used in amateur bands,

the parasitic elements are spaced from the radiator by 0.15 to 0.25 wavelength and are about five percent longer or shorter than a half-wavelength, depending upon the function of the parasitic and the array spacing. Directivity is through the shortest parasitic element.

The number of parasitic elements also enters into the determination of optimum parasitic length, power gain, directivity, and bandwidth. Needless to say, as the operating frequency of the Yagi array is varied from the design frequency by a few percent, the parasitic elements become detuned from optimum, and over-all antenna gain drops sharply, especially when the parasitic elements become self-resonant. In some instances, depending on parasitic element tuning, the directional radiation pattern is obscured, or even reversed, with maximum gain occurring in the reverse direction of the array.

Antenna power gain and beam directivity degeneration commonly occur when the parasitic element tuning is incorrect. Thus, the frequency span of the Yagi antenna must be restricted to that narrow region over which the parasitic elements remain in proper phase relationship. To increase the bandwidth of the Yagi, the parasitic elements must be detuned from optimum, which decreases power gain and reduces front-to-back ratio. The gain-bandwidth product of the Yagi must therefore be sacrificed to some extent to obtain an increase in either factor. In any case, the product is small but is a critical factor of adjustment. To increase the bandwidth without a corresponding decrease in gain, log-periodic principles may be applied to the Yagi antenna.

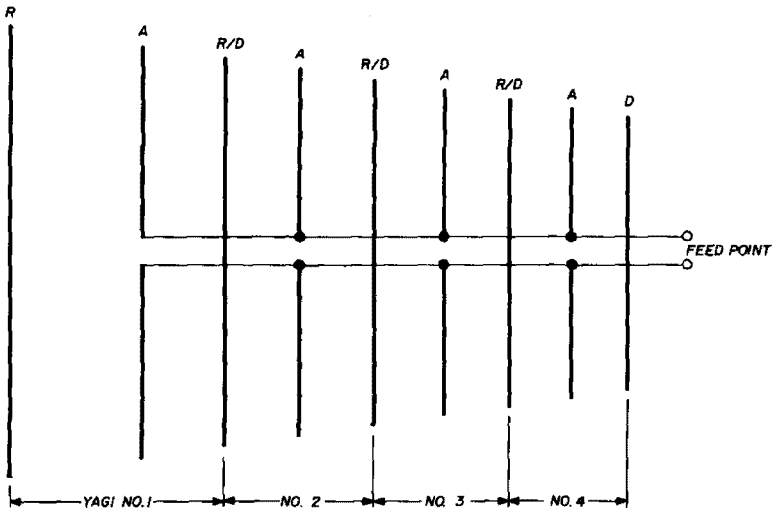


fig. 4. Log-periodic Yagi performance is a result of Yagi arrays with parasitic elements serving dual function of director and reflector for adjacent larger and smaller Yagi arrays.

the log-periodic yagi

A log-periodic Yagi (LPY) array may be constructed of individual Yagi antennas differing in size by a geometric constant, properly arranged and fed. A simple LPY antenna is shown in **fig. 4.** The LPY is made of a series of end-fire Yagis, with each driven element fed from a common balanced transmission line. Unlike the driven element in an LPD antenna, those of the LPY are fed in a non-transposed manner. The in-between elements are parasitics, and log-periodic performance is obtained by making each parasitic element serve the dual function of director and reflector for the adjacent larger and smaller driven elements.

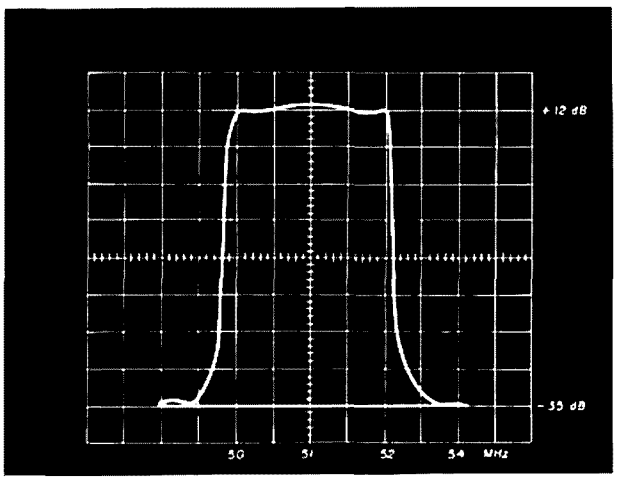
Practical LPY antennas with power gains of about 9 dB have been built for the 1.1- to 1.25-GHz range.[5] Over-all length of the LPY antenna is large for the gain produced, especially considering the high power gain per Yagi array normally obtained for a given size.

the bandpass antenna

An interesting and practical variation of the basic LPY antenna is the **LPY bandpass array**, which provides greater power gain per unit length. This unique antenna makes use of a log-periodic dipole struc-

ture having the frequency characteristic of a Chebyschev-type filter. A number of parasitic director elements, trimmed to cover the appropriate frequency range are used to enhance the power gain of the log-periodic array. A frequency sweep of such an antenna designed for the six-meter band is shown in the photograph. The LPY bandpass antenna is easy to build, simple to adjust, and provides good power gain considering the over-all length of the structure. The original LPY bandpass design was evolved for long-distance color-tv reception, which demanded a combination of good passband characteristics, high gain, and good adjacent channel discrimination. The LPY bandpass antenna combines these attributes, and various versions of this antenna are now used for tv and amateur two and six meter band operation. They are commercially available.*

An effective six-meter band LPY bandpass antenna is shown in **fig. 5** and the photo. It's composed of five log-periodic elements and three parasitic directors. The antenna has been in use for some months at K6RIL and at other California six-meter stations. Results compare favorably with an eight-element Yagi on a thirty-foot boom. Gain is estimated at 12 dB or better; front-to-back ratio is apparently about 24 dB. The design range is 50 to 52 MHz.



Gain-vs-frequency characteristic of the six-meter LPY antenna.

Response on reception is down about 20 dB at 47 and 53 MHz. It provides some receiver protection from spill-over from nearby channel-2 tv transmitters, because antenna response is down a comfortable number of decibels at frequencies higher than 54 MHz. Antenna response is down 20 dB at 45 degrees either side of center at 51 MHz. This is representative of the pattern over the antenna's operational range.

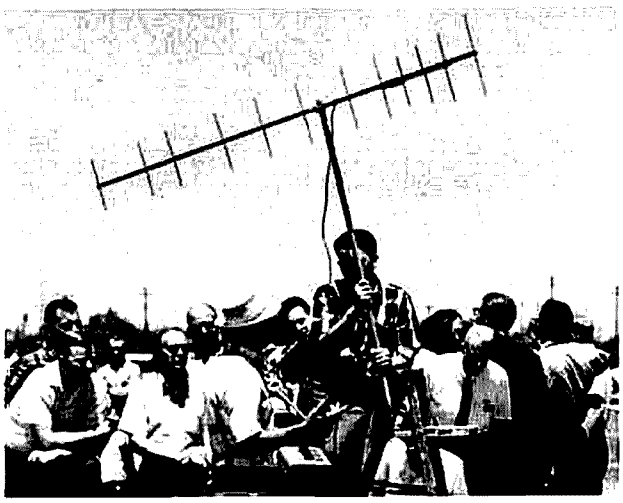


photo by W6BUR

Oliver Swan with a wideband 12-element 400 to 450 MHz antenna at the West Coast VHF Conference antenna-measuring contest; gain measured at approximately 7.5 dB.

Input impedance is about 75 ohms. The antenna can be fed with a 70-ohm coax line (RG-11A/U) and a balun, or a matching device can be placed at the antenna apex for use with a 50-ohm coax line or high-impedance balanced line. The line at K6RIL is 75-ohm heavy-duty tv ribbon.

construction

The commercial version of this antenna is built on a double boom, much in the manner of heavy-duty tv antennas (see

* LPY bandpass antennas were originally designed by Oliver Swan and are for sale by Swan Antenna Company, 646 North Union Street, Stockton, California 95205.

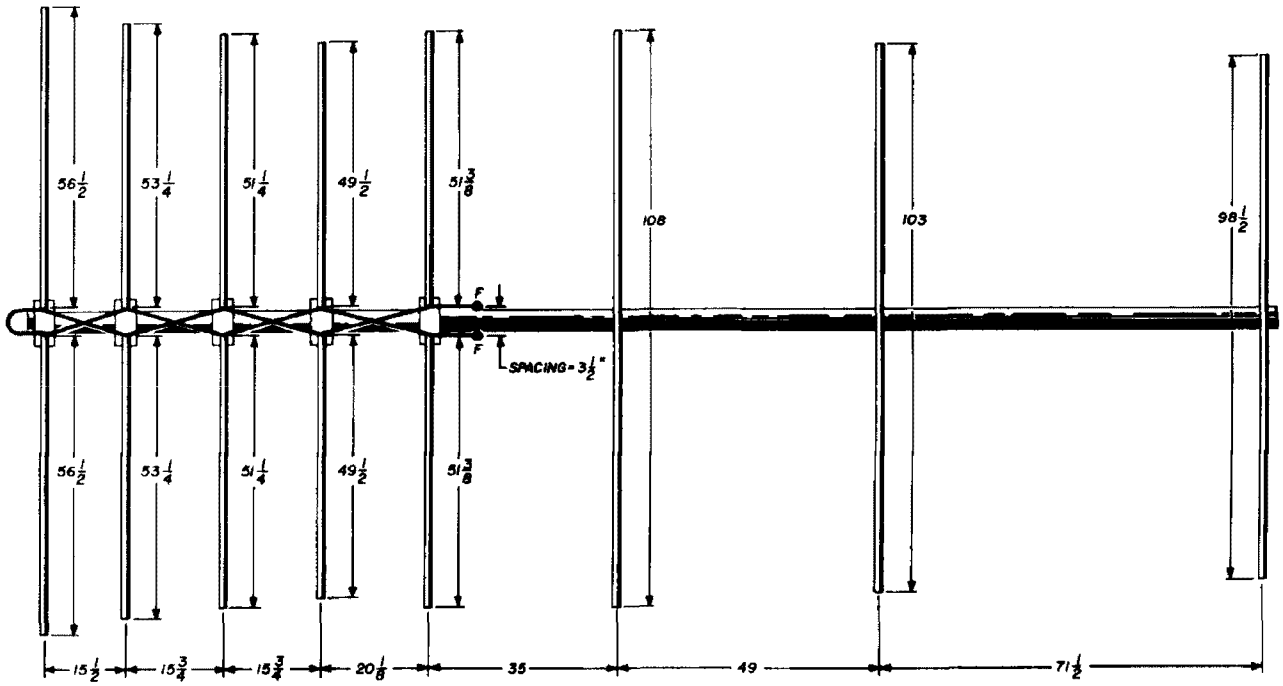


fig. 5. The LPY bandpass antenna. Dimensions (in inches) are for 6 meters (50 to 52 MHz) although they may be scaled for other frequency ranges. Elements are supported above a 2-inch aluminum boom. Half lengths are given for the log-periodic elements and full lengths for the parasitic elements.

photograph). A simpler configuration for those wishing to build their own array would be the use of sections of 2-inch aluminum tubing for the boom, as shown in **fig. 5**. The driven elements are supported at the center on insulating blocks, and the parasitic elements are mounted directly to the metal boom. Since a simple log-periodic structure is used with no interspersed parasitics, the driven elements are cross-connected with aluminum clothesline wire. The last driven element to the rear is shunted with a six-inch loop of wire at the end of the transposed feedline.

A balanced feedpoint exists at the front of the log-periodic assembly, and a balun or matching device may be mounted at this point. The array is supported from the vertical mast structure at the array's center of gravity. An overhead support to each end of the boom is recommended.

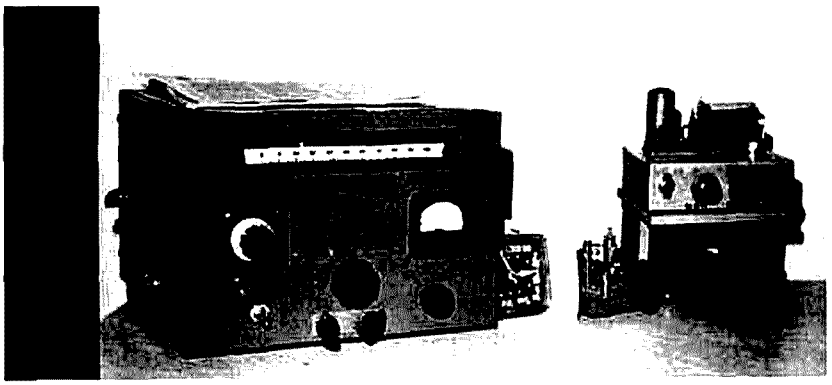
operation

The LPY bandpass antenna has been in operation at K6RIL for some months and is used in conjunction with an eight-element Yagi on a thirty-foot boom, mounted on a nearby tower. Numerous tests on transmission and reception have shown that the two antennas are nearly identical in performance as far as gain and front-to-back ratio are concerned, within the bandpass of the Yagi antenna. Operation of the LPY bandpass antenna at the extremes of the passband shows superior performance compared to the Yagi. The LPY bandpass array will be the "antenna to watch" on two and six meters in the coming months.

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2. Smith, "Log-Periodic Antenna Design Handbook," Smith Electronics Company, Cleveland, Ohio.
3. Duttamel and Isbell, "Broadband Logarithmically Periodic Antenna Structures," 1957 IRE National Convention Record, Part One.
4. Isbell, "Log-Periodic Dipole Arrays," University of Illinois Antenna Laboratory Technical Report 39, June, 1959.
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ham radio



cw transceiver for 40 and 80 meters

The best features of transistors and tubes are combined in this complete station designed for portable use

Clifford Bader, W3NNL, 1209 Gateway Lane, West Chester, Pennsylvania 19380
Richard Klinman, K3OIO, 1339 Pennwood Road, Philadelphia, Pennsylvania 19151

A casual inspection of current amateur radio publications reveals an interesting fact: the huge rack-and-panel stations of fifteen or twenty years ago are gone. In their place are complete radio stations that occupy just about the same volume as a 1950 communications receiver. Today is the era of the transceiver, which is hard to beat for compactness and real operating convenience.

The transceiver described in these pages resulted from a desire for a good portable station for trips, hunting expeditions and contest work. A rather unorthodox circuit is used. The usual multiple-heterodyne circuits are eliminated together with their elaborate shielding systems. "Straight-through" transmitter circuits are featured, and the received signal is immediately converted to audio frequencies. The design objectives were:

1. Operation on the 40- and 80-meter cw bands.

2. Electronic and mechanical ruggedness.

3. Simplicity; a single vfo for receiver and transmitter.

4. Operation for at least twelve hours on a single 12-volt auto battery. The receiver must operate from a dry battery if necessary.

5. Power inputs from ten to thirty watts, preferably higher; but with good efficiency.

design considerations

Meeting these objectives isn't easy, especially if you're restricted by time and economic considerations. The problem was solved by using inexpensive bipolar transistors and junction fet's in all circuits except the transmitter driver and output stages. In these stages the old reliable 12AU6 and 1625 were used. This may seem to be a throwback to 1945, but it proved to be an excellent choice. The 1625 is practically indestructible, draws only 450 mA heater current, is inexpensive, and works well with low plate voltages. Also, with no modification, a higher plate voltage on the 1625 tube allows up to 75 watts input if energy consumption is not important.

On the other hand, rf power transistors for 40 meters are costly, were not immediately available, and would require more than 12 volts, so that no saving in power supplies would be obtained. Furthermore, it would be difficult to achieve efficient operation at 10 and 75 watts input; also, the risk of transistor burnout would be present when tired operators weren't careful in tuning up or in making sure the antenna still existed. The delicate nature of rf power transistors is widely noted in the literature.

This article is intended to present the amateur community with some new ideas and to stimulate the application of ingenuity and effort.

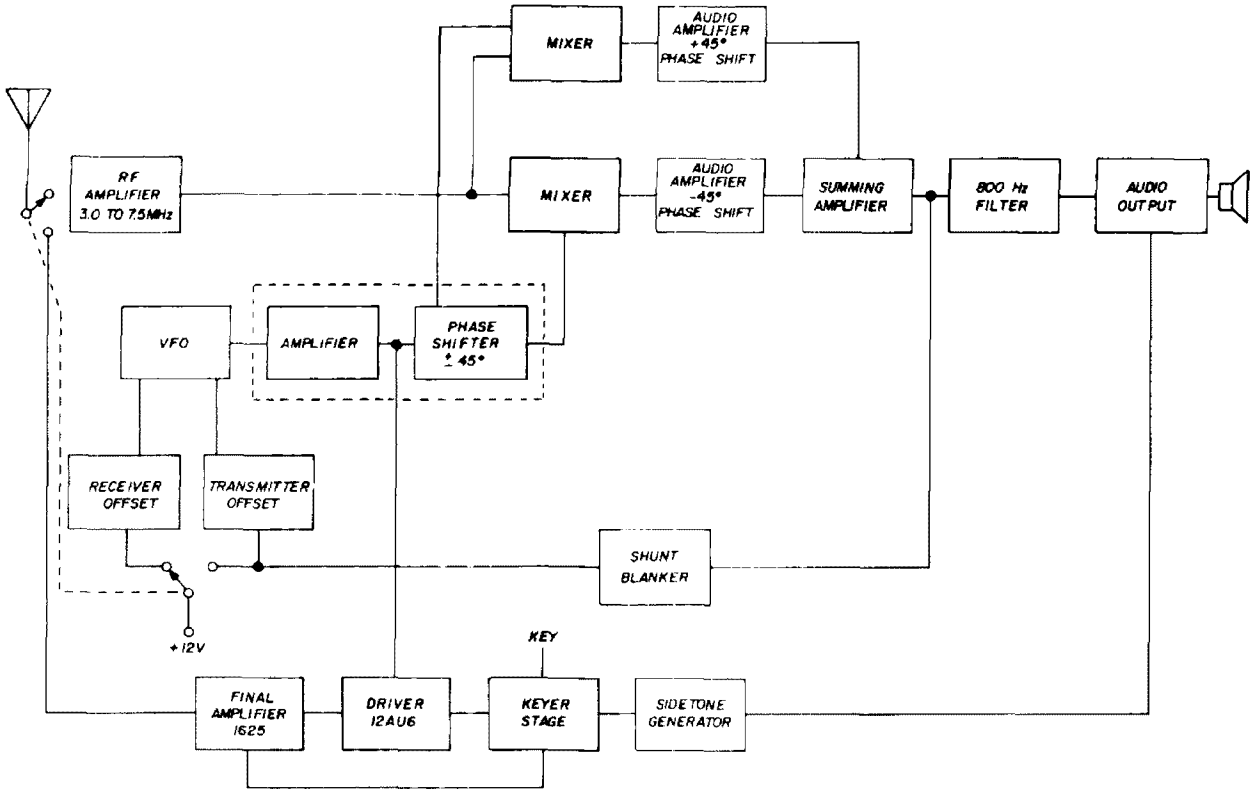


fig. 1. Block diagram of the CW transceiver for 40 and 80 meters.

results

The block diagram of **fig. 1** shows the receiver as simply a trf circuit with a product detector.[1] The same vfo and frequency multipliers are used for both transmitter and receiver. For receiving, all that's needed is an audio amplifier, which is easily constructed with transistors. Audio filtering at 800 Hz provides selectivity. The audio image is suppressed by a phasing system that reduces

bandwidth and provides single-signal reception. The undesired sideband remains at least 10 dB down (calculated) at 1600 and 400 Hz.

The transmit-receive function is accomplished by a toggle switch in the interest of simplicity. Cathode keying is used, and transistor buffering removes high voltage and current from the key. An audio oscillator is used for a transmitting monitor.

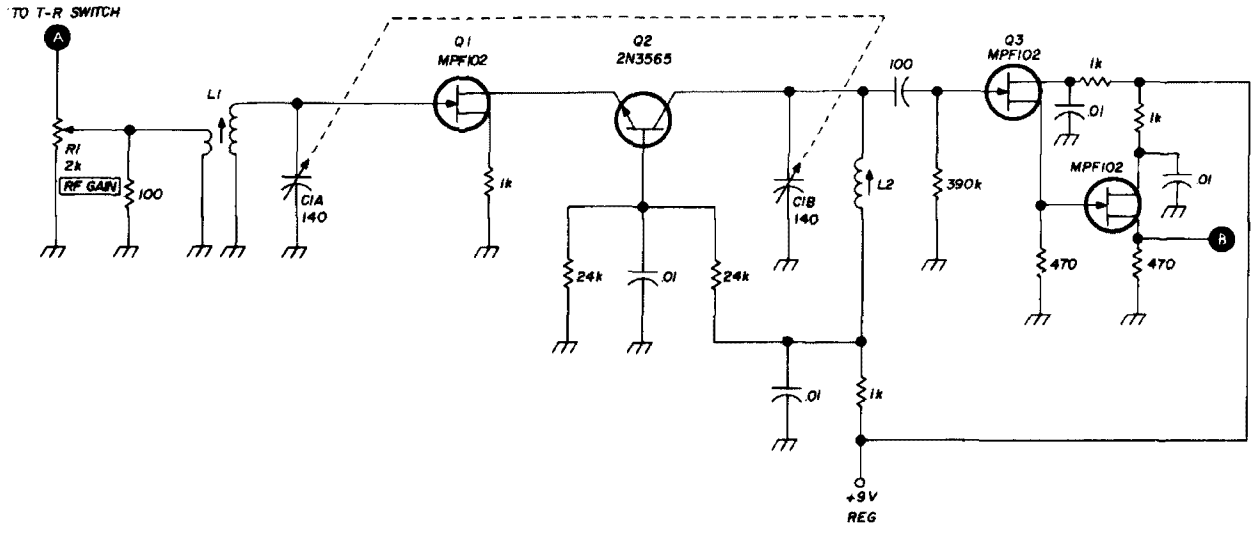
the receiving section

The receiver is shown in **figs. 2** through 7. Only a small amount of rf amplification is needed at these frequencies, and most of the signal processing and amplification is done at audio frequencies.

Since it is difficult to control the gain of transistor stages while maintaining optimum intermodulation characteristics, the rf-gain control, R1, is a simple attenuator in the antenna circuit. The cascode rf amplifier uses an N-channel junction fet, Q1, followed by a common-base buffer stage, Q2. Both input and output tanks, L1, C1 and L2, are high-Q tuned circuits. High Q is very important here, since this is the only place where unwanted strong signals are rejected and cross-modulation effects minimized. Tuning of these tanks is fairly critical. C1, a front panel control, must be repeaked at least every 50 kHz (during receive only) for optimum results. The circuits tune from 3 to 7.5 MHz.

the vfo

The vfo **(fig. 4)** consists of a jfet oscillator, Q8, and buffer, Q9, in a Hartley circuit operating between 1.75 and 1.85 MHz. No heat is present near the oscillator tank, so the vfo is extremely stable. The rugged construction of L3 and C2 provides good mechanical stability.



C1 Dual 140-pF variable (Hammarlund HFD-140)

L1, L2 37 turns no. 27 enamelled, close wound on a Millan 69046 slug-tuned form; 3-turn link

fig. 2. The rf amplifier uses a cascade circuit with an fet followed by a common-base buffer.

In the receive position of the t-r switch, S1, air trimmer C3 is inserted into the tank section by reed relay K1. Alternatively, K2 selects trimmer C4 during transmit. These trimmers provide receiver offset (incremental tuning). Incremental tuning is required after calling CQ or during a contact, because any tuning with the main tuning capacitor, C2, will also change transmitter frequency.

exciters

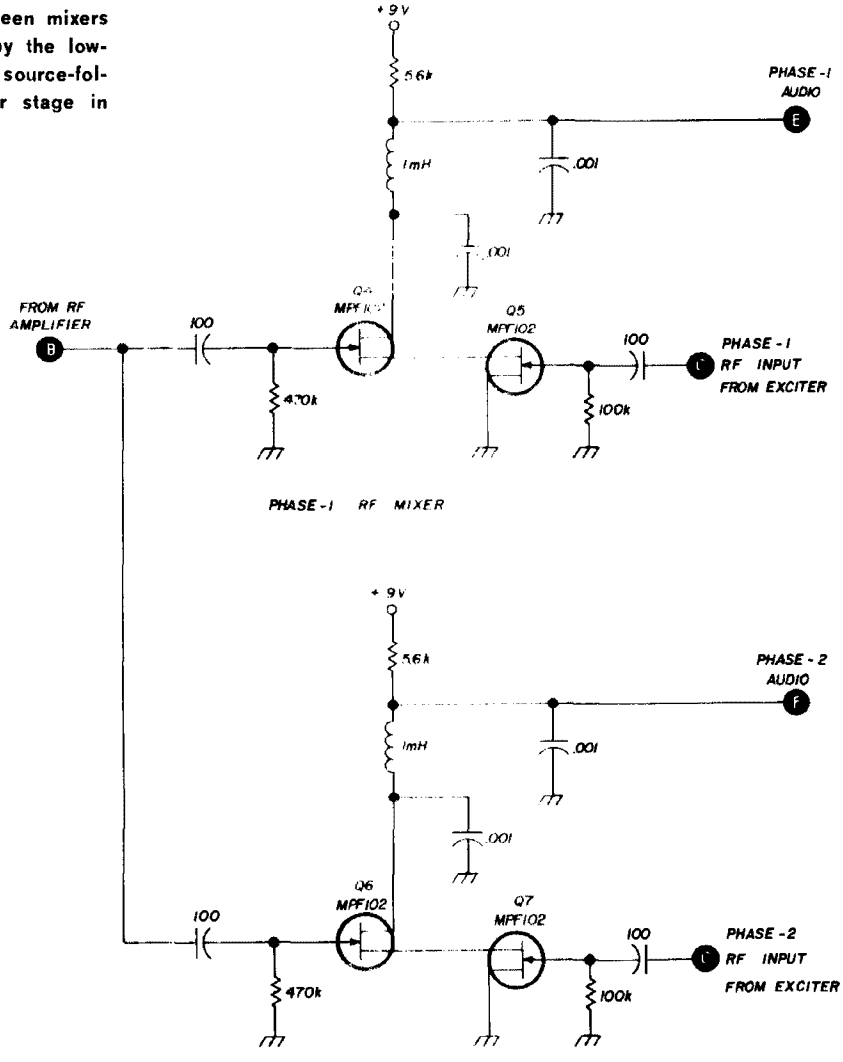
Two plug-in jfet exciters are used, one per band for 80 and 40. (See **fig.** 5 and 6.) These provide transmitter drive voltage and the

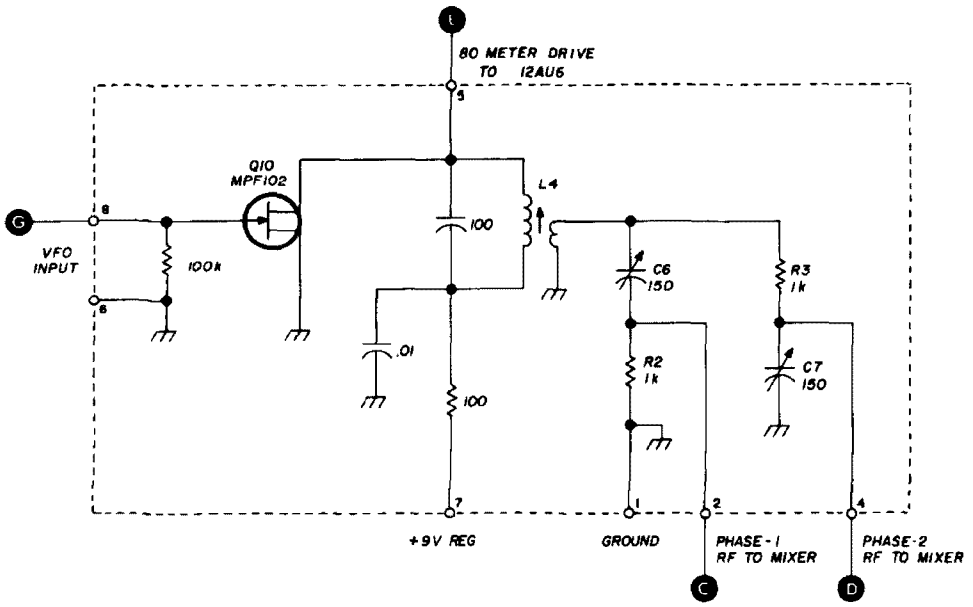
±45-degree phase shift for the vfo signal during receive. The driver output, Q10 or Q12, is fed to the 12AU6 grid and to RC phase-shift networks, C6, C7, R2, R3; or C9, C10, R4, R5, which yield two 80- or 40-meter voltages 90 degrees out of phase. Tuned circuit L4 or L5 ensures the proper harmonic of the vfo.

A jfet source follower buffer, Q3, provides a low-impedance source and prevents interaction between the series mixers. The same rf signal is mixed with each of the vfo outputs, yielding two audio signals 90-degrees out of phase at 800 Hz. The audio signals are preamplified by Q13, Q14, Q15 and Q16. One channel has a +45-degree phase shift at 800 Hz provided by C11 and R8; the other has a −45-degree phase shift provided by R9 and C12. The result is a second 90-degree phase shift difference added to the quadrature audio signals. The phased audio signals are summed in amplifier Q17-Q18, and one sideband is cancelled. Prior to summing, the signals are adjusted to identical amplitudes by balance control R7.

Depending on the way the rf and audio phasings are compounded either upper or lower sideband may be selected for single-

fig. 3. Series mixer stages for the receiver section. Interaction between mixers is limited by the low-impedance source-follower buffer stage in fig. 2.





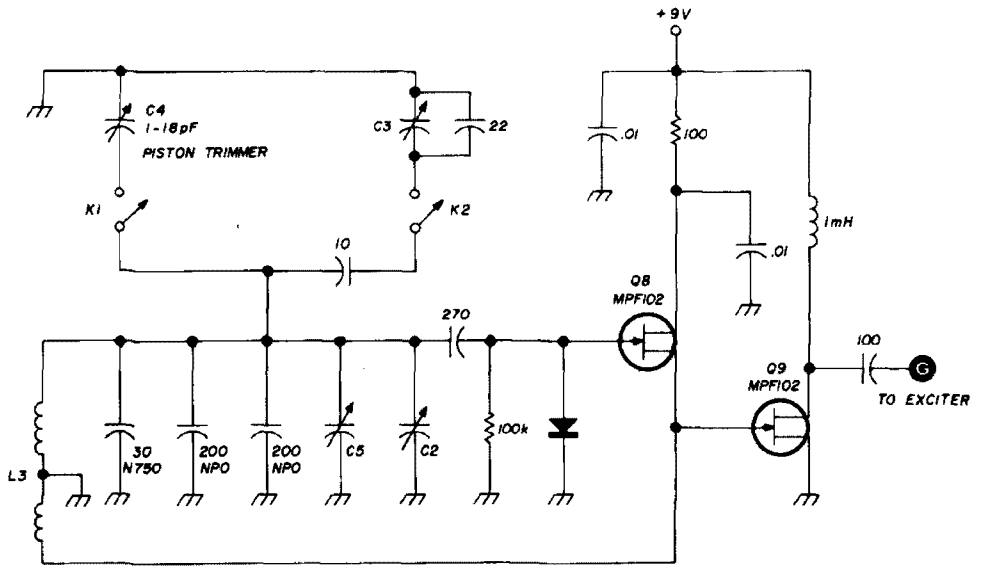
L4 45 turns no. 36 enamelled close wound on Milien 69058 form, slug-tuned; link is 25 turns no. 36

C6, C7 150-pF double mica padders (El Menco type 424)

fig. 5. Plug-in exciter unit for 80 meters.

signal reception. The unit currently tunes **above** zero beat.

The audio signal is amplified in an operational amplifier, with a bridged-T selective filter at 800 Hz in the feedback loop. Gain is high through the amplifier only at the fil-



C2 100-pF double-bearing variable

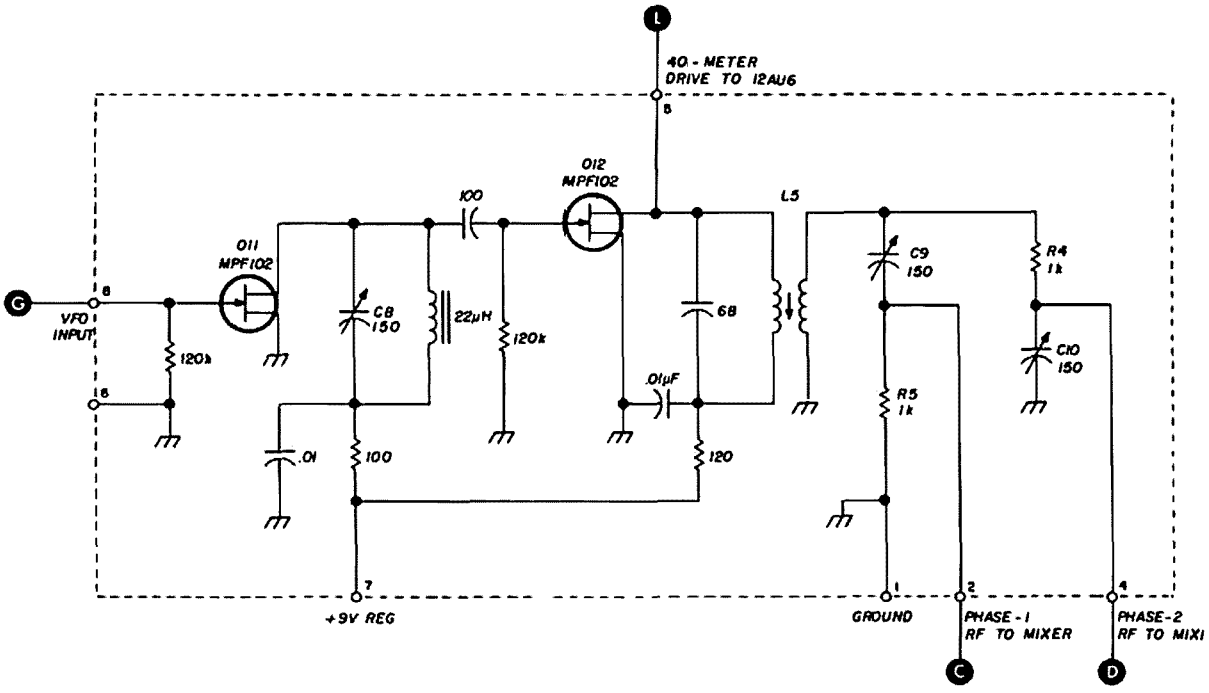
C3 Hammarlund APC-35 with all but three stator and three rotor plates removed

C4 18-pF piston trimmers (JFD VC32GWY)

C5 50-pF variable (APC-50-screw adjustment only)

L3 24 turns no. 24 enamelled on 1 1/4" diameter form, wound to occupy 1 1/4" with 7-turn link of no. 24 enamelled interwound at ground end of main coil

fig. 4. The vfo uses a Hartley circuit that tunes from 1.75 to 1.85 MHz.



C8, C9, C10 150-pF double mica padders (El Monco type 424)

L5 30 turns no. 36 enamelled close wound on Millen 69058 form, slug-tuned; link is 10 turns no. 36

fig. 6. Plug-in exciter for 40 meters.

ter frequency (**fig. 7**). Selectivity is adequate for cw, and sufficient skirt response makes tuning uncritical. The receiver is blanked during transmissions by shunt gate Q19, which is across the gain control, R6.

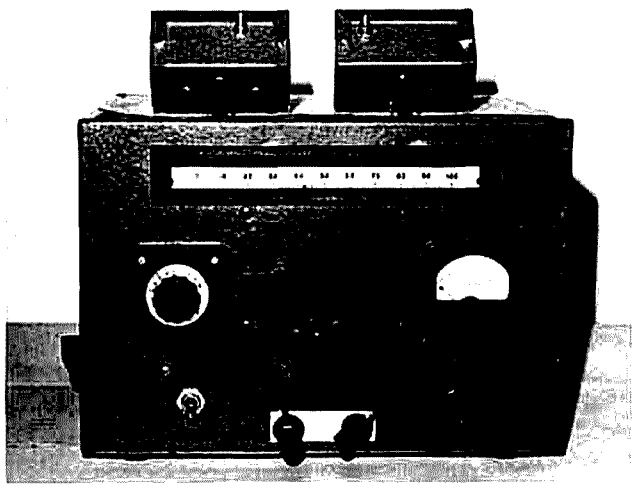
Operational amplifier Q23, Q25, Q26, Q27 and Q28 provides power amplification. A transformerless complementary output stage is provided by Q26 and Q27. The circuit is conventional except for the fet constant-current source composed of Q28 and Q29. This circuit provides adequate drive to Q27 during negative-signal excursions, while minimizing the current drain on Q25 during positive signal excursions.

Either the received signal or the sidetone oscillator output of Q37 is fed to Q23 base, which is the summing input. The output stage will drive loads as low as eight ohms, although a higher impedance is preferable. A 25-ohm speaker is normally used, but high impedance headphones can be substituted.

The only remaining receiver circuit is a 9-volt regulator for stable vfo and rf operation, **fig. 8**. The reference is the reverse-biased, base-emitter junction of Q30 acting in the breakdown mode. The regulator is a feedback type using difference amplifier Q31-Q32 to drive a series regulator, Q33.

transmitter section

The transmitter is conventional (**fig. 9**). The exciters, which are the same units used for receiving, provide frequency multiplica-



The 80- and 40-meter transceiver. Plug-in exciter units are on top.

tion and drive for straight-through operation of the driver and final amplifier tubes. The 12AU6 provides more drive with fewer stages than could be obtained with transistors. Drive is controlled by a potentiometer in the 12AU6 screen grid. The final grid tank tunes 3.5 to 7.5 MHz without switching. The final is capacitance-bridge neutralized. Its pi-network output tank will match a wide range of input impedances. The taps and capacitors, switched by S2, will match high or low tube impedances to reasonably flat coax.

keyer, monitor and t-r switch

Both tubes are keyed simultaneously by Q34, Q35 and Q36. Cathode keying eliminates the standby current drain of the 12AU6. For the 12AU6, an ordinary transistor (Q36) that can take more than 50 volts at low current is used as a keyer; an inexpensive RCA 40327 (Q35) is used for the final keyer. The 1625 cathode contains a shaping circuit composed of C16, C17, D17, R12, R13 and R14. The diode-resistor network prevents the capacitors from discharging too rapidly through the transistor on "make." The keying wave shape on "break" is affected by the 12AU6 cathode bypass capacitor. A conventional metering circuit is used to measure final cathode current.

Audio oscillator Q37 provides a sidetone for monitoring. This is a simple unijunction relaxation oscillator powered from the same transistor (Q34) that gates and provides current to cathode switches Q36 and Q35.

The t-r switch is a dpdt toggle switch. One contact transfers the antenna; the other selects the proper vfo shift relay and supplies the keyer with power during transmit.

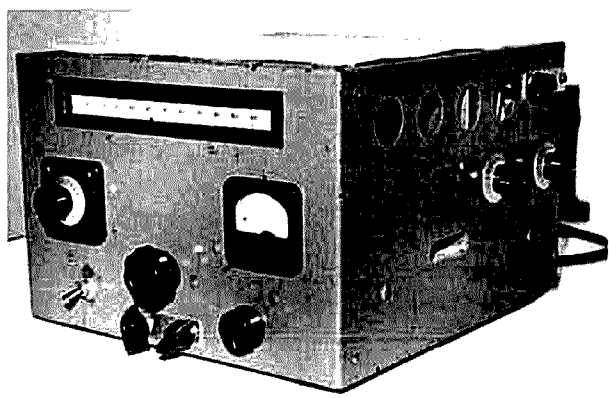
power supply

The power supply shown in fig. **10** is the one used for fixed station operation. It is presented in the interest of completeness, but of course, any conventional supply could be used that delivers the voltages shown. Surplus transformers can be used, and the diodes are available from many surplus sources listed in the amateur literature. The values for all components were chosen with conservative ratings to ensure reliability for portable use.

construction

Circuit layout isn't critical, although good construction practice should be followed to ensure stable and reliable operation. Parts substitution is recommended, and a well-stocked junk box would be an asset.

The transceiver was built into the case and chassis of an old National FB7 receiver because it was available and was the right size (8-inches high x 12-inches deep x 12-inches wide). It happened to contain an excellent 160-meter vfo built by Earle Lewis, W3JKX. In fact, the vfo was used exactly as it was with the exception of a 6SJ7, which was replaced by jfet's Q8 and Q9. A few resistors were changed, and the supply voltages were adjusted for the transistors.



Side view showing the pi-network controls. Grid tuning and drive level controls are on the other side.

There were plenty of holes to mount circuit boards and controls. Most of the transistor circuits were built on printed circuit cards containing a matrix of rectangular pads, but tie strips or made-to-order printed circuits could be used. The exciters are built in 3¼-inch x 2-inch x 1½-inch miniboxes fitted with chassis-mounting octal plugs.

controls

The front panel controls are main tuning, which is the original receiver dial with a linear logging scale; receiver incremental tuning; rf and af gain controls; t-r switch; and a final-amplifier cathode-current meter (200 mA full scale). Other less-used controls are located where convenient. The set-and-

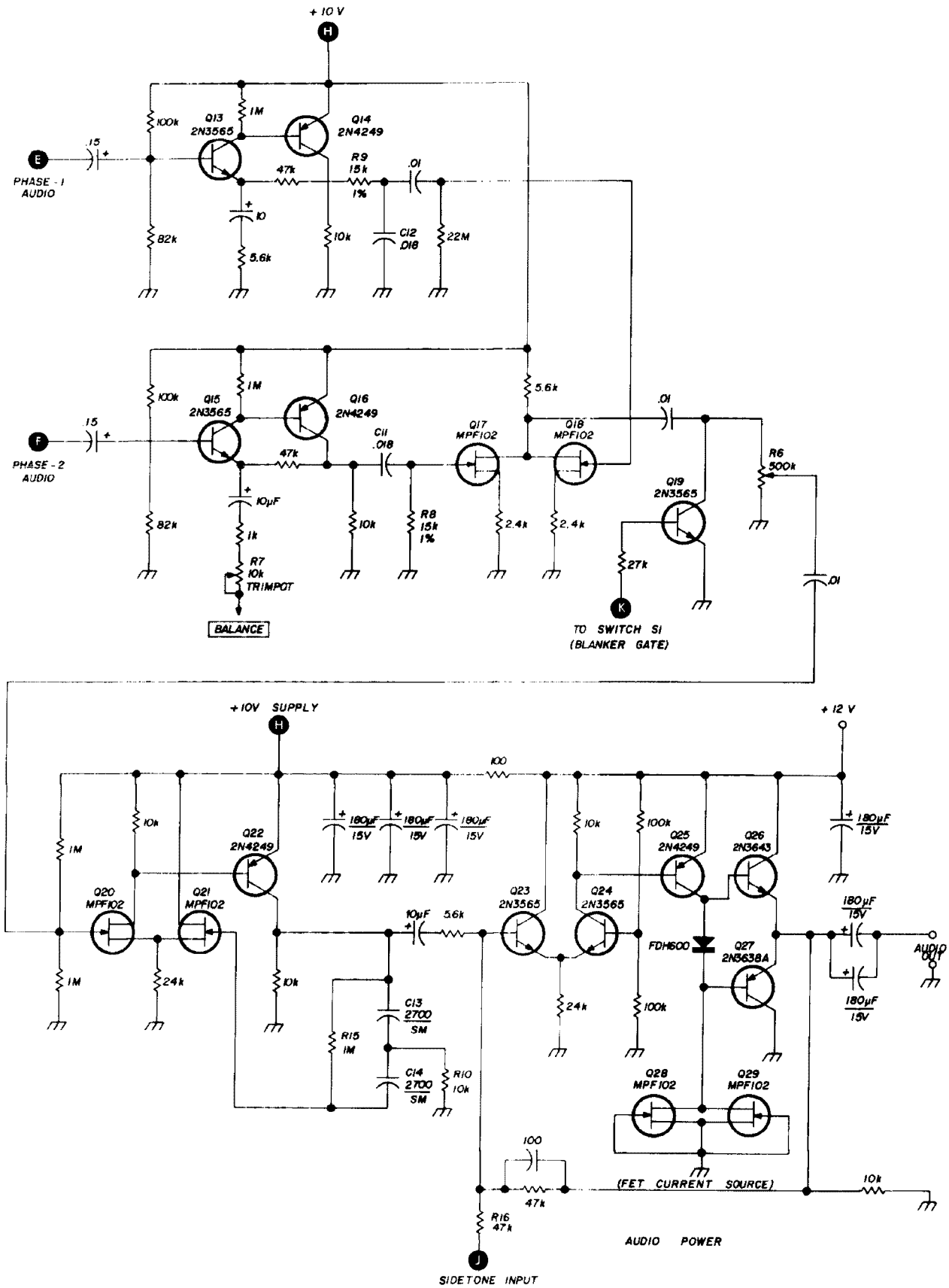


fig. 7. Audio section includas a summing amplifier, blanker gate, 800-Hz filter and audio power stage. R7, a 10-turn pot, should be grounded.

forget neutralization capacitor, C21, is above the torpedo-mounted 1625. The rf plug-in exciters are accessible through the top lid. On one side of the cabinet are the final grid tuning, C15, and drive-level control, R11. On the other side of the case are the final amplifier pi-network components, C18, L7, S2 and C19. The audio output, key jacks, and antenna connector are on the rear wall. Power is brought in by an eight-wire cable terminated with an octal male connector.

As we said before, construction is not critical as long as good layout procedure is followed. A few noteworthy points are in order though. The receiver 12-volt supply and the transmitter filament grounds are fed through separate wires. This keeps tube filament currents and corresponding IR drops from appearing as modulation on the receiver supply. During ac operation, this voltage may cause hum.

Physical layout of the phase-shift networks should be symmetrical to provide identical electrical operation. Also, short rf and ground leads are essential.

Transistor circuits should be isolated from the final amplifier tank. In the unit shown, the t-r switch is on the front panel, but is enclosed in an aluminum box for isolation. The reed relays are mounted next to the vfo tank circuit. Short leads of heavy wire are used for mechanical and electrical stability.

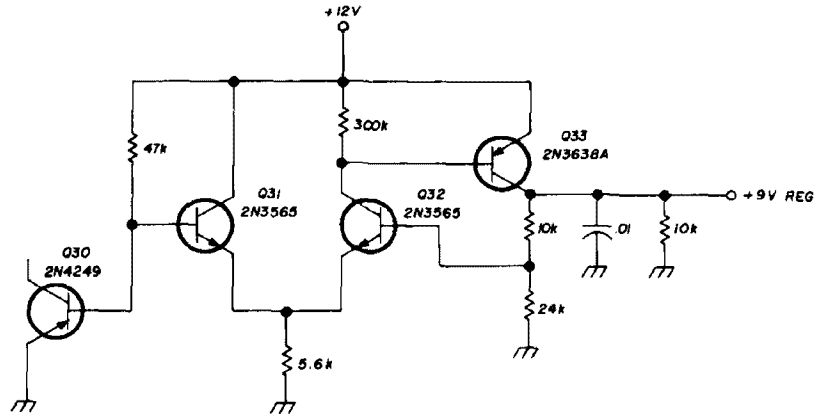
audio and sidetone frequencies

If you don't particularly like an 800-Hz note, filter C13, R10, C14 and R15 can be tuned to another frequency. Any good handbook will give values for a bridged-T network. Sidetone volume can be changed by adjusting R16 (**fig. 7**); higher values will reduce audio level. Sidetone frequency can be changed by varying C20, R17 or both (**fig. 9**).

receiver section adjustment

You'll need an auxiliary receiver to calibrate and adjust the transceiver. With application of receiver voltage only, the vfo can be set by adjusting C15 while monitoring the signal on another receiver. Only a logging scale is used, so desired bandspread is ensured by choice of vfo tank LC ratio.

fig. 8. Nine-volt regulator circuit for stable operation of the vfo and rf operation.



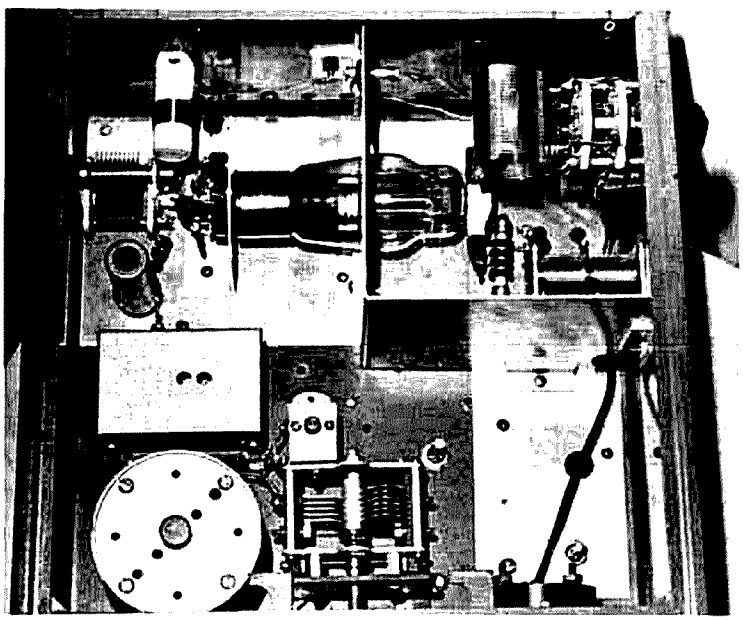
The rf section must be peaked to the desired band after the exciter is plugged in. This is done by setting C1 and varying L1-L2 for maximum received signal. Signals on both bands are peaked by setting C1 and varying L1-L2. L4 (3.5 MHz) and L5 (7 MHz) should be tuned for maximum signal.

After setting main tuning capacitor C2 on the high side of a received signal to obtain an 800-Hz note, or at the audio frequency peaked by the audio filter, the signal is peaked by carefully setting C6-C7 on 3.5 MHz and C9-C10 on 7 MHz. The phasing is now approximately correct, but the audio channel gains must be equalized by adjusting balance control R7. The receiver is now tuned to the same beat frequency on the undesired sideband, and R7 is adjusted until the best null is obtained, which should also correspond to a good signal on the proper

sideband. Finally, C6-C7 or C9-C10 may be retrimmed for greatest null depth on the undesired sideband.

The final part of the receiver adjustment is the frequency offset. Incremental tuning control C3 should be set near midrange. While listening to the vfo on a receiver, C4 is adjusted for no change in vfo frequency when switching S1 from receive to transmit. This means that the receiver and transmitter are in zero beat. Incremental tuning is provided with a numbered knob.

After the offset is set to zero beat and the position of C3 recorded, C3 is set to receive



Internal layout of the transceiver. Power amplifier and pi network are in the rear; vfo and plug-in exciter are to the front.

a note of approximately 800 Hz on the proper (high) side of zero beat as determined by peak-audio response. This normal setting is recorded for future reference and will be different for each band, while the zero offset point will be the same.

transmitter section adjustment

The first step is to neutralize the amplifier, using C21, according to the procedure outlined in the ARRL handbook.

Fig. 9 can be used for setting the pi-network tap positions. After moving the t-r switch to transmit, the sidetone note should be audible when the key is closed, even with no high voltages. With plate and screen voltages present, cathode current should also flow when the key is closed.

Advance the drive control, R11, to maximum. Set C15 near maximum capacity for 80 meters and near minimum capacity for 40 meters. Quickly dip the final using C18, with C19 at maximum capacity. Load the final to 50 mA cathode current (with 750 volts on the plate), then back off on the drive, R11, until an increase in plate current is just barely noticed. Tune the final grid for minimum plate current. The final may now be loaded to the input power desired.

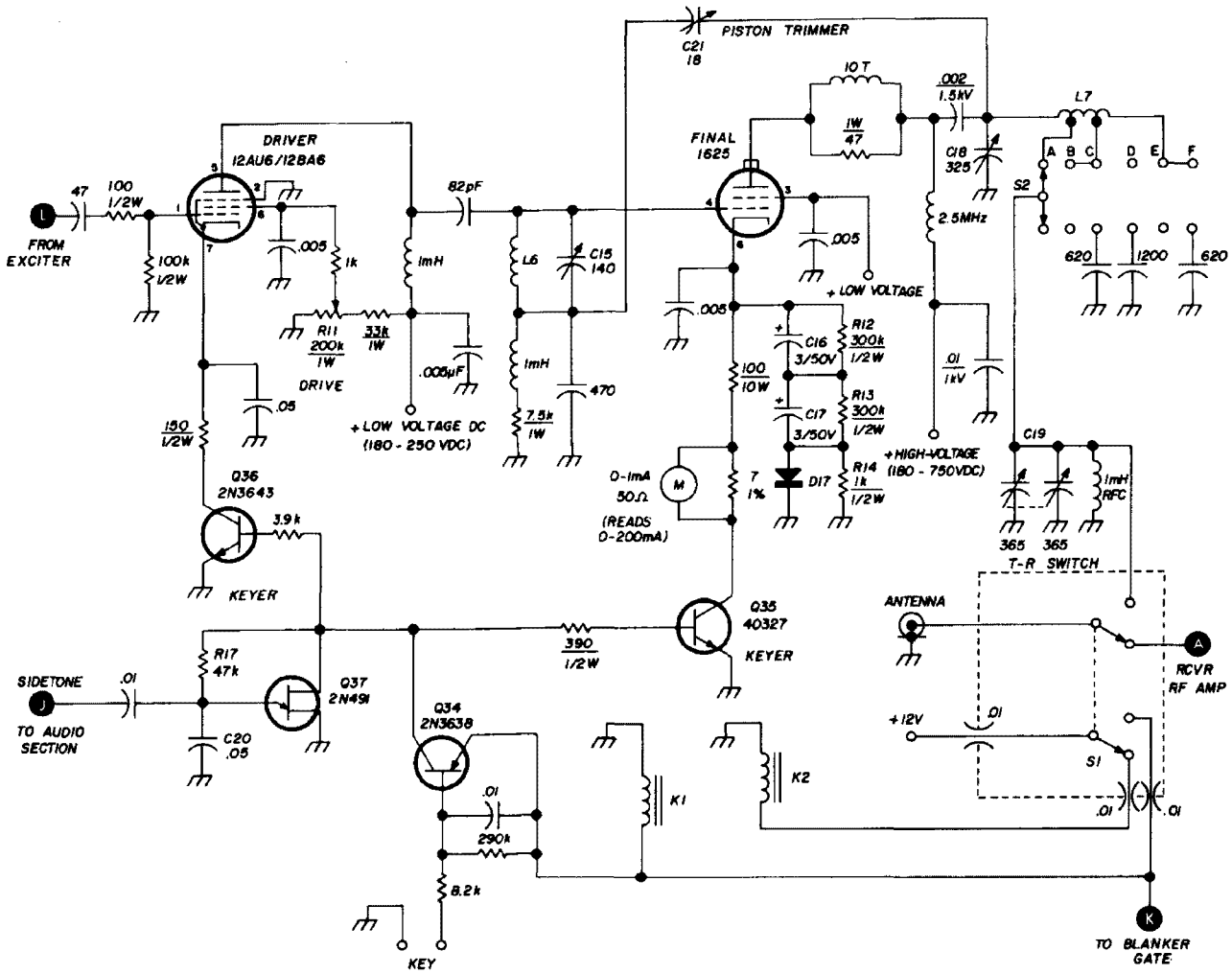
operation

It was a pleasant surprise to find that in no case did any transistor stage react unfavorably to the full 75 watts input. With full input the vfo remains exactly on frequency, and the transistor keying stages don't latch up. In fact the rig is now used with no ground, a random-length antenna is fed through an LC tuner, and performance is unaffected.

During transmit the vfo determines the frequency, but in receive the frequency may also be changed by the incremental tuning control, C3. When answering or calling stations the received signal is peaked at an 800 Hz note, while the offset tuning is set at the predetermined "normal" point.

Since C3 and C4 have been adjusted for an 800 Hz offset between received and transmitted frequencies, you can be reasonably sure of being in zero beat with the desired station. Precise zero beat is obtained by adjusting the receiver offset tuning to the predetermined zero offset point where the received and transmitted frequencies coincide. Zeroing the received signal with the vfo then ensures zero beat. The offset tuning may then be adjusted for the best signal, which should be at about the normal setting.

After contact is made, the receiver should be tuned with the offset control, C3, because changing the main tuning will change transmitter frequency.



C15	**140-pF air variable (Hammarlund MC 140)**
C18	**325-pF air variable (Hammarlund MC 325)**
C19	**dual 365-pF (broadcast receiver type)**
L6	**27 turns no. 25 enamelled close wound on 3/4" diameter form**
L7	**30 turns B&W 3019, tapped at 11 turns (A) and 19 turns (tap B-C) from plate end of coil**
K1, K2	**Miniature reed relay, 200-ohm coil (Wheelock 3421)**
S1	**dpdt toggle**
S2	**6-position, 2-pole, rotary ceramic switch**

fig. 9. Transmitter driver, final and keyer circuits. Switch S2 different ranges of antenna and vacuum-tube impedance: A, 40 meters, low tube impedance; B, 40 meters, high tube impedance; C, 80 meters, low tube impedance; D and E, 80 meters, high tube impedance.

progress report

The transceiver has been operating since June, 1968. During the ARRL field day contest, nearly 300 contacts were made, including some with the West Coast, using 10 watts input.

At present, K3010 is using the transceiver as a permanent station. It runs 50 watts input and feeds a 100-foot wire antenna with excellent results. Much DX is heard and worked; the best to date is Europe on 80 meters and Australia on 40. You can hear the rig on the 80-meter Eastern Pennsylvania Traffic Net. The transceiver is ideal for this purpose because it's ready to go with no warm up.

improvements

Among the improvements that would be desirable are break-in operation and in-

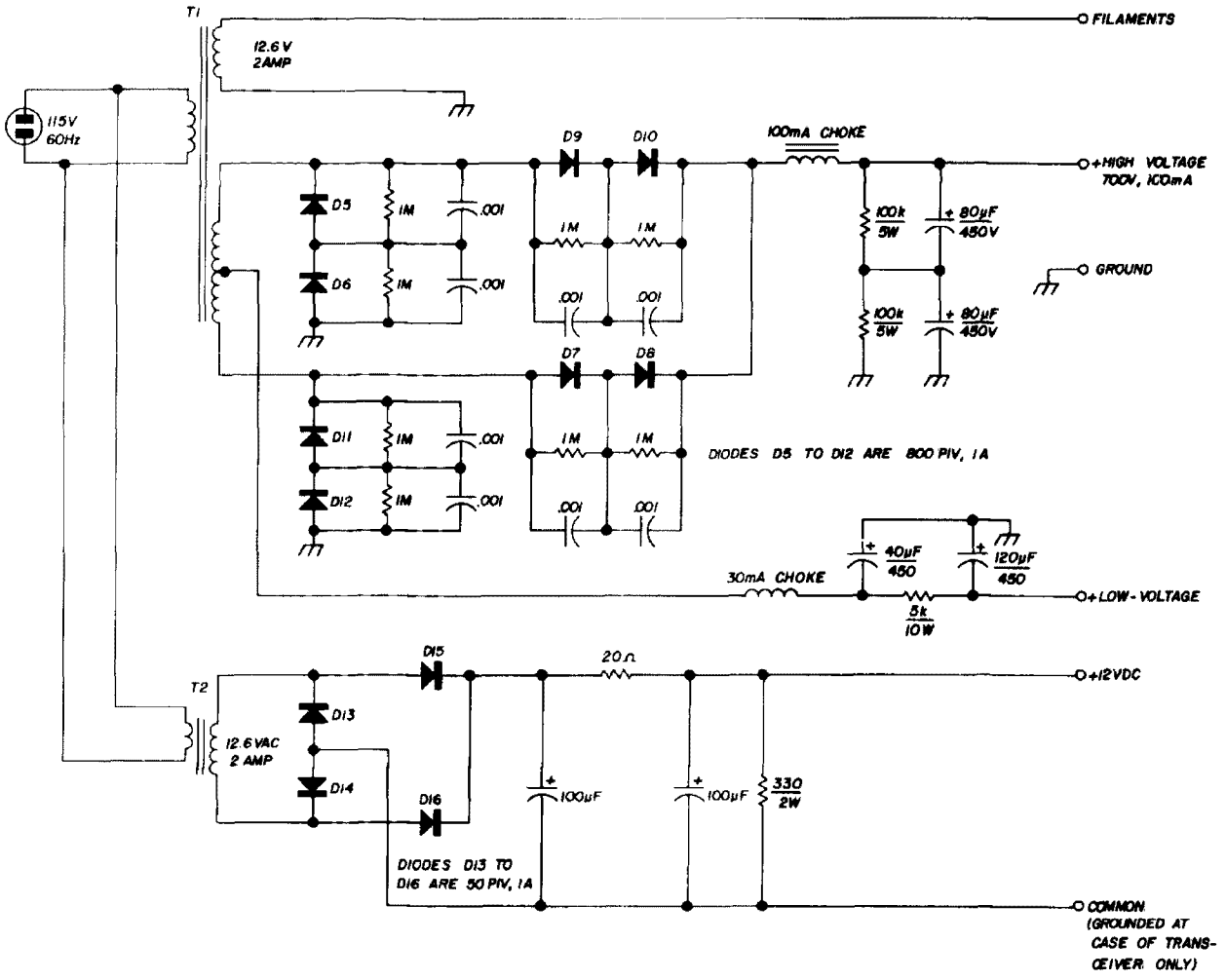


fig. 10. Fixed-station power supply which is suitable for this transceiver. Transformer T1 is an old tv power transformer; T2 is a 12-volt filament transformer.

creased selectivity and sensitivity. Also, cross-modulation could be reduced by optimizing the circuits. The mixers should have good intermodulation characteristics, since any beats between adjacent strong signals in the rf passband will appear in the audio output. Also appearing in the audio output will be modulation on signals strong enough to cause significant variation of mixer operating point due to a limited dynamic range.

The receiver is less tolerant in this respect than a superheterodyne, since many beats generated by nonlinearity in the first mixer of the latter will fall outside the i-f passband. The best arrangement would appear to be a balanced mixer in which device nonlinearities tend to cancel and dynamic range is increased. In addition, an increase in local-oscillator drive amplitude would be helpful. Such measures might be necessary, for example, if the transceiver were to be used in a field-day setup with adjacent 80- and 75-meter stations.

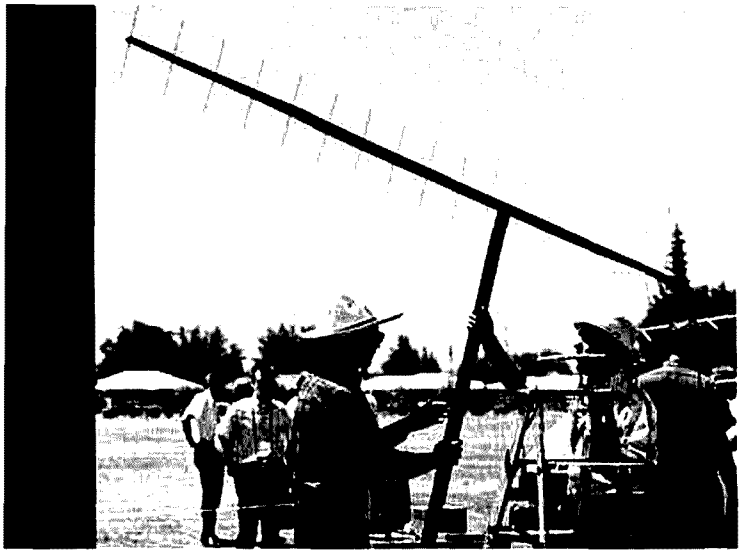
ssb operation

The receiver does a good job on single sideband. However, sideband suppression for transmitting would require a wideband audio phase-shift network. The audio-frequency amplifiers and wideband network could be switched between transmitter and receiver sections. With this and linear operation of the final amplifier, the transceiver could be used on single sideband. Anyone interested in this challenging project is invited to suggest ideas.

reference

1. White, "Balanced Detector in a TRF Receiver," *QST*, May, 1961.

ham radio



direct methods for measuring antenna gain

How to obtain meaningful data using simple equipment

Bruce Clark, K6IYO, RdF Corporation, Western Regional Office, Suite 401, 19702 South Main Street, Gardena, California 90247

For the amateur interested in top station performance on any band, antenna refinement definitely produces the most rewarding return per unit of effort and expense.

Only in the antenna system, which includes the feedline and supporting structure as well as the radiator, can improvements increase performance for both transmitting and receiving. Unfortunately, however, the antenna system is usually the most neglected part of an amateur station. Performance tuning, if done at all, is usually limited to adjusting the driven element length, sliding the clamps on the T match, or adjusting the gamma capacitor for the lowest standing wave ratio. Except for using the swr bridge, "antenna scope" impedance bridge or field-strength meter, most amateurs seem content to leave antenna tuning to the manufacturers.

The manufacturers can't build antennas to meet all performance demands. Commercially built antennas are designed for "average" installation conditions. All too often these just don't exist in most amateur installations. Most amateurs are plagued by poor soil conductivity, height restrictions, nearby objects

and a host of other adverse conditions that affect antenna performance. These adverse effects can be reduced by tuning the antenna system once you have some dependable quantitative data as a baseline for optimization.

The degree of improvement by tuning is limited with simple antennas. With the more elaborate arrays used above 14 MHz, it's possible to obtain performance increases up to 3 dB with small antennas. Improvements of 7 to 8 dB are possible with larger arrays.

The following paragraphs present simple methods for measuring vhf antenna gain directly, with good accuracy. Once you **know** what your antenna is doing, you can make the right adjustments to optimize performance. A few examples are also given of some rather startling results obtained by amateurs who were introduced to these methods.

direct measurements

The average amateur can measure antenna gain with adequate precision using simple equipment. The measurement results are much more meaningful than, say, a measured standing-wave ratio of 1.02-to-1 on the transmission line. All this indicates is that the antenna is taking power. It may or may not be radiating in the desired direction or with the desired efficiency.

Of the many methods of measuring antenna gain, two are within the capability of the amateur. These are (a) the attenuator/receiver method, and (b) the matched-detector method. Both are comparison tests using a reference antenna and the antenna to be tested. Received signals provide the measurement data.

These methods are more reliable and provide more repeatable data under varying site conditions than those using transmitted signal/field-strength meter or measured-pattern methods.

attenuator/receiver method

The attenuator/receiver method is block diagrammed in **fig. 1.** Basically, the system uses an accurate attenuator combined with the station receiving system. The signal-source output should be as low as possible and still provide a usable signal at the receiver S-meter when the reference antenna is connected. For most situations 100 mW is adequate. The source should be stable and free of spurious outputs.

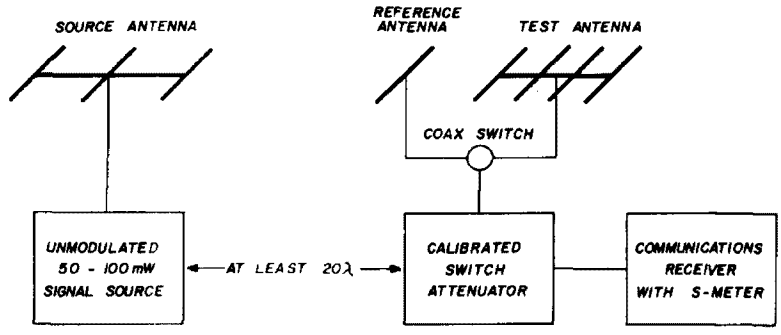


fig. 1. Test equipment for the attenuator/receiver method. The source antenna should be as high as possible, in the clear, and at least 20 wavelengths from the antenna under test.

procedure

Set up the source antenna in the clear at least 20 wavelengths from the test antenna. A nearby amateur's tower, flag pole or tv mast is a good support. Turn on the source, and adjust the attenuator for a reference level on the receiver (anywhere between S-6 and S-9 will do). Record the number of dB used on the attenuator to obtain the reference value on the S meter. Switch to the test antenna, and peak the antenna for maximum signal. Adjust the attenuator for the same S-meter reading obtained with the reference antenna. Record the new attenuator reading. The difference between attenuator readings is the amount of gain (or loss) between the two antennas.

Repeat the process several times, moving the reference antenna for an average level. Note that some variation is introduced by moving the reference antenna. This can be reduced by using a directional source antenna to reduce ground reflection contributions to the received signal (discussed later). In addition, the source antenna should be moved between several different sites at varying distances. Several measurements should be made at each site. The resultant gain figure should be the average of at least six readings.

Note also that feedline losses are included in these measurements. If known, they can be added to the measured antenna gain to get the actual gain of the antenna. Although less impressive, the measured figure is a more practical value, especially above 50 MHz where feedline loss contributions are significant.

The attenuator/receiver method will give accuracies on the order of ±1 dB. It's limited by the accuracy and resolution of the attenuators, but is probably the most applicable method for amateur work.

matched detector method

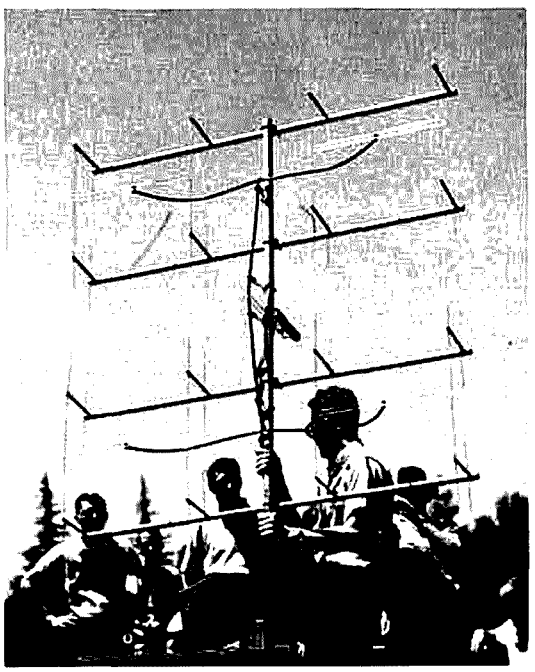
The matched detector method is very popular with the West-Coast vhf crew. It requires more sophisticated equipment, but gives greater resolution and quicker readout. The absolute accuracy is still limited by the reference antenna performance due to reflections. The averaging procedure should be used here also if absolute gain figures are desired.

Either a high (1-W) or low (10-mW) source signal, modulated with a 1-kHz audio tone is used (**fig. 2**). The type of source determines the detector type. A crystal diode detector similar to a Telonic XD-series, or a homebrew equivalent[1], will give a square-law output at low input levels. This is ideal for the vswr meter readout.

procedure

The vswr meter is a 1-kHz, sharply tuned, gain-stable, low-noise audio amplifier driving an rms ac vtvm. The 1-kHz modulation is detected and amplified. This signal drives the meter, which is calibrated directly in dB. By adjusting the vswr meter gain range (0-60 dB in 10-dB steps), a reference level can be obtained with the reference antenna. The test antenna is then connected to the detector, and the gain increase (or decrease) noted.

Although the initial cost is high ($200), the vswr meter is available in surplus outlets for approximately $40 to $60 for the earlier Hewlett-Packard HP-415 series. Others by PRD and General Radio are also available.



K6JYO's extended 32-element collinear provides 15 dB gain.

If the source power is too high, the crystal diode detector will be driven out of the nonlinear, square-law portion of its curve. The resultant output will deviate, and the vswr meter reading will be high. This can be prevented by:

1. Keeping the source power output very low.

2. Inserting a calibrated attenuator (3 to 6 dB) ahead of the detector mount (**fig. 2**).

3. Using the vswr meter with a wider range detector called a bolometer (thermistor) mount. Although not as sensitive, the bolometer mount provides good results when used with sources of 1 to 2 watts output.

the reference antenna

No study of antenna gain measurement would be complete without a word on reference antennas. Classically, the isotropic radiator is a point source that illuminates all points equally on the inside surface of a sphere. It is used as the reference antenna in nearly all theoretical work. However, it's not possible to produce such an antenna, so the matched 1/2-wave dipole[2] is used as a reference antenna.

The dipole has a disadvantage. Because of its broad pattern, it's extremely sensitive to ground effects and to near-field reflections from the signal source. These reflections add or detract from the desired free-space signal and produce an output that varies from the ideal (average) value.

nbs standard reference antenna

Absolute accuracy of measurements depends on the accuracy of the reference dipole. Hence it is important to average the reference dipole readings under different site conditions. Recently, highly accurate standard reference antennas have been designed and employed by the National Bureau of Standards (NBS) and some amateurs, among them W6VSV and W6HPH. Basically a simple directional array designed for low side-lobe content and high front-to-back ratio, the NBS standard antenna has a gain of 7.7 dB over a reference dipole, measured under laboratory conditions in an anechoic chamber.

The measurement repeatability is on the order of ±0.1 dB or better. The NBS standard antenna is used in a manner identical to that of the reference dipole, but there is less variation due to reflections. Also, one must remember to add the 7.7 dB reference-antenna gain figure to those from the vswr meter with the test antenna in the line. For example, if the test antenna measures 2.3 dB when the reference antenna measures 0 dB, the antenna gain is 10 dB.

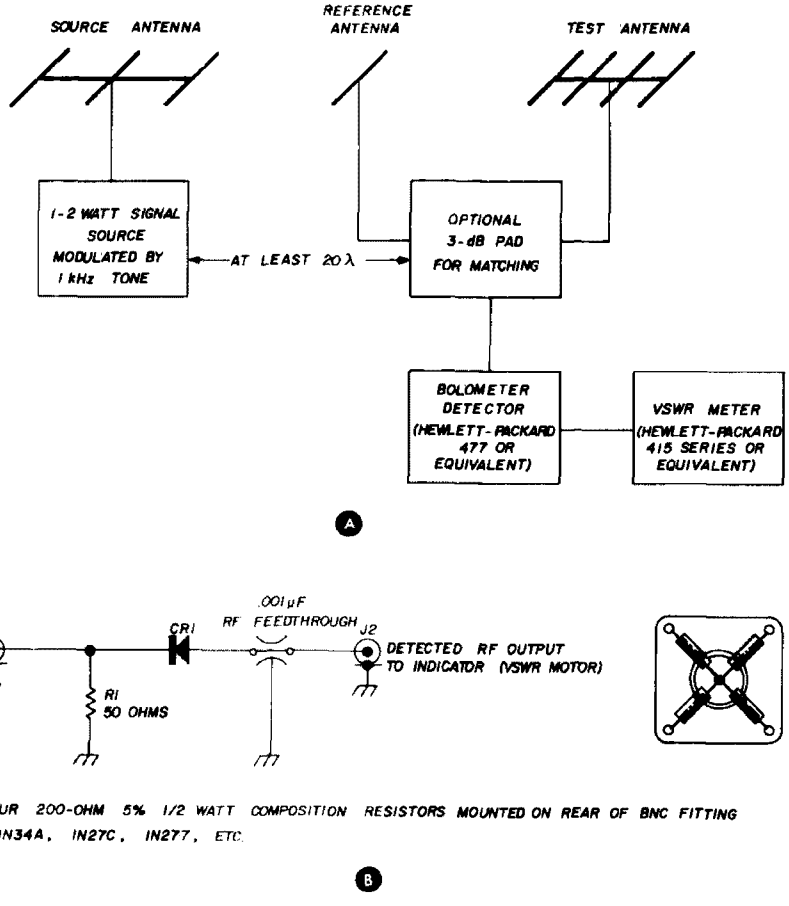


fig. 2. Setup for the matched detector method. A bolometer is used in (A). An easily constructed diode detector substitute for the bolometer is shown in (B). The 3-dB pad will improve the match between detector and antenna, especially at uhf.

results

These techniques are regularly employed by top vhf-uhf amateurs to obtain the most from homebrew and commercial arrays. In the past few years, antenna contests at hamfests have become popular proving grounds where new winning combinations have been discovered. A case in point is the reawakened popularity of the Yagi antenna at 432 MHz. It has resulted from careful optimization of several scaled-down designs that didn't work at all (or poorly at best). Another case is the 1 to 2 dB gain increase from adding directors to collinear arrays—a method now adopted by at least one manufacturer.

The accuracy of the results is amazing. My own 32-element, 432-MHz array measured 15 dB at the West-Coast Uhf Conference in Fresno and 16.2 dB at the Hughes Radio Club contest in Fullerton (after some matching deficiencies were discovered).

As for the repeatability of results from site-to-site, tests of the popular 6-foot boom "Tilton" Yagi at 432 MHz resulted in consistent measurements yielding 12 to 13 dB in contests from Missouri to California. W5ORH's twin bi-square beam measured 8.0 dB at three different sites using three different test methods. These examples are exceptions. Typically, however, results haven't varied more than ±2 dB when good equipment and normal care were used in making the measurements.

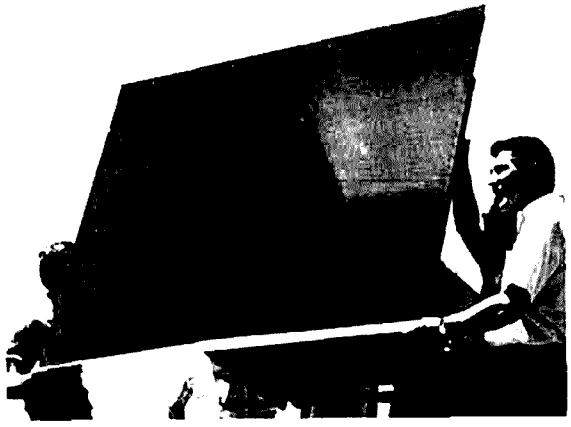
W6FZJ and his 13.3-dB 1296-MHz array.



WA6KKK and WB6MGZ aim 18.6-dB 1296-MHz dish.



W6MMU's big horn for 432—11.7 dB.



K6MYC's 12-element 432-MHz quad gave 7.9 dB.

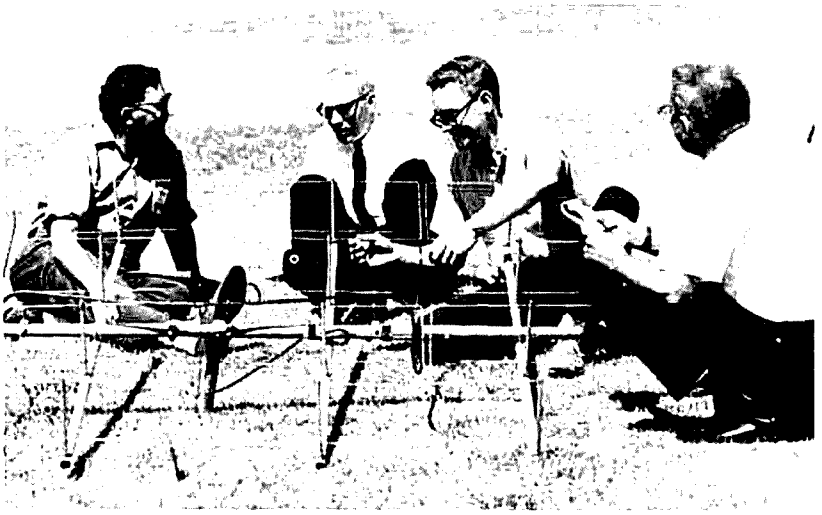






K7ICW's 30-element Yagi for 1296 MHz yielded —2.5 dB!

VK3ATN, W1DTY, K6JYO and W6DOR discuss some of the intricacies of antennas for 432 MHz while attending the antenna measuring contest held at the West Coast VHF Conference in Fresno.



photos by W6BUR

some surprises

At one contest several owners of supposedly high-gain commercial arrays really had their eyes opened. One 432-MHz Yagi, with a manufacturer's claim of "over 17 dB forward gain," measured **negative** 2 dB off the front and +6 dB off the back. Cutting the antenna in half got about +8 dB forward gain.

Another homebrew 13-element Yagi from a popular vhf handbook measured +1.9 dB gain over a dipole. (The owner had substituted a wooden boom for the original metal boom and hadn't reduced the element lengths to compensate. Trimming the elements and matching the feed brought the gain up to 12.3 dB—not a bad increase.)

It should be obvious that antenna gain measurement is worthwhile for the amateur. From my experience, it gets results we all desire: better reports and more consistent contacts.

reference

1. Fred W. Brown, W6HPH, "The Matched Detector," *VHF'er*, June, 1965.
2. Loren Parks, K7AAD, "The Reference Dipole," *VHF'er*, April, 1965.

ham radio

the crystal oscillator

This complete summary of solid-state devices as crystal oscillators will enhance your technical reference file

Hank Olson, W6GXN, P. O. Box 339, Menlo Park, California 94025

In a previous article,[1] I discussed the general nature of crystal oscillators using tubes, transistors and field-effect transistors (fet's). Since 1966, when that article was written, a great deal has happened in the semiconductor industry that might affect your choice of an active device for use in a crystal oscillator. In the following paragraphs, I've considered all the solid-state devices available at reasonable cost that can be used in crystal oscillator service.

A brief review of their application in conventional circuits is first presented. This is followed by an extensive treatment of these units as used in modified versions of the basic circuits. I've also given some recommendations for certain precautions and design considerations that should be used. If you're solid-state oriented, this article will be invaluable as a reference source the next time you consider a crystal oscillator design.

Prices have dropped, and performance has increased significantly on n-channel junction fet's, insulated gate fet's, digital integrated circuits and linear integrated circuits. The price reductions appear to be due mainly to lower-cost plastic packaging and the acceptance of these devices by large commercial makers of computers and television sets. The lower costs and the increased performance are just what the doctor ordered for amateur applications.

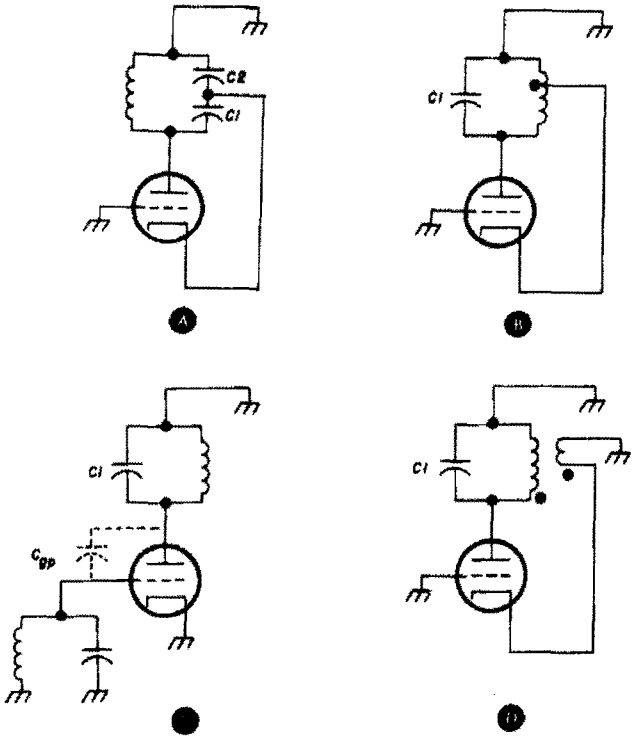
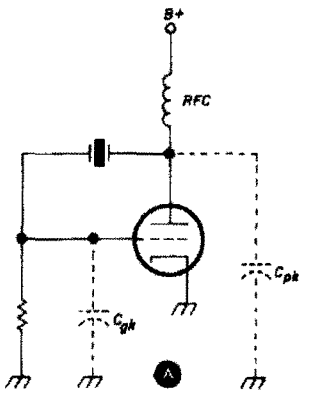


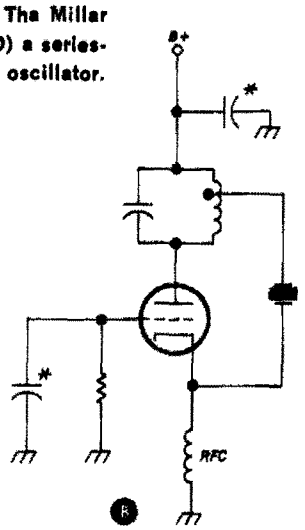
fig. 1. The four basic oscillator circuits: Colpitts (A), Hartley (B), tuned plate-tuned grid (C) and tickler-feedback (D).

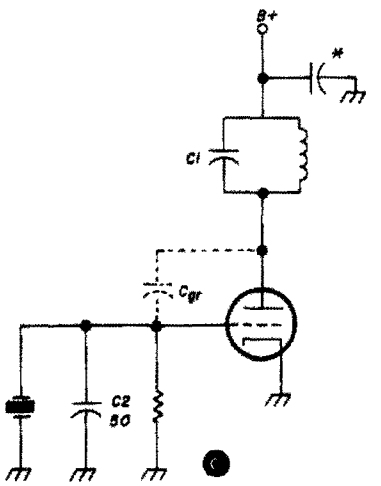
basic circuits

Four forms of the vacuum-tube oscillator are generally used in amateur designs. These are the Colpitts, Hartley, tuned-plate-tuned-grid and tickler-feedback circuits, as shown in **fig. 1**. They can all be modified to incorporate a crystal, as shown in **fig. 2**. Note that some of these circuits use the crystal as a series-resonant circuit, and others as an inductance. A single crystal can present either of these characteristics, as shown in **fig. 3;** but you must account for the fact that **each occurs at a different frequency.** This is one reason for the various "CR" specifications; some crystals are cut for series-resonant and some for near parallel-resonant operation. A listing of military CR specifications is given in **table 1**.

fig. 2. Descendants of the oscillators in fig. 1 using crystals for frequency control. In the Pierce (A) Cpk and Cgk replace C and C2 of fig. 1(A). Series-mode crystal is used in the Hartley circuit (B). The Miller oscillator (C) is a version of the tptg. In (D) a series-mode crystal is used in a tickler-feedback oscillator. Blocking capacitors are denoted by *.







modified forms

One of the most successful of the vacuum-tube circuits, for use with series-resonant crystals, is shown in **fig. 4**, the Butler oscillator. It's similar to that in **fig.**

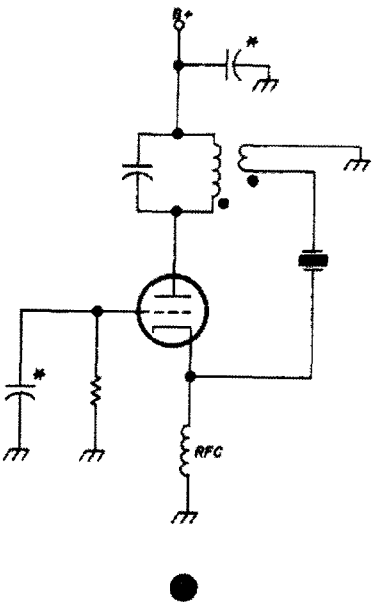
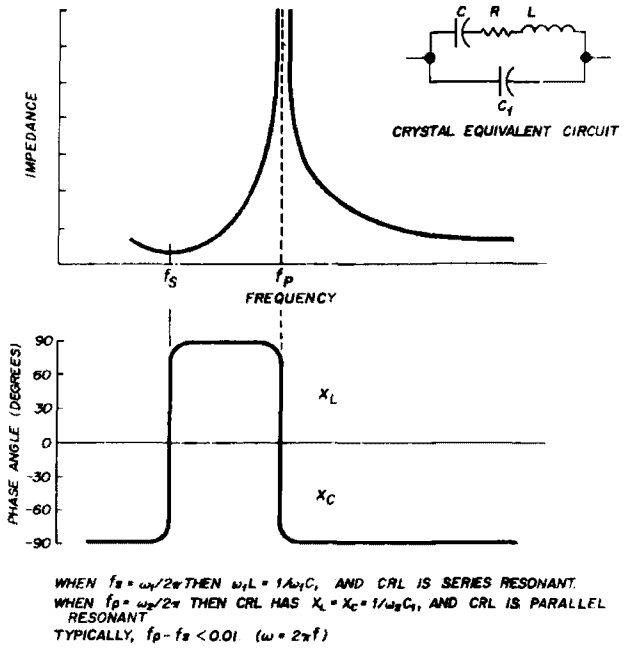


fig. 3. Series and parallel resonance impedance characteristics of a crystal.



1A, except that an impedance-lowering cathode-follower has been added, so that the low impedances of the series-resonant crystal and cathode of V_1 can be more easily driven.

Two other types of crystal oscillators somewhat familiar to hams are the negative-resistance oscillator, as exemplified by the transitron and dynatron oscillators using tubes, and the tunnel-diode oscillator of the solid-state world. The general negative-resistance crystal oscillator is shown in **fig. 5,** and the transitron and dynatron in **fig. 6.** Two tunnel-diode crystal oscillators are shown in **fig. 7.**

From **fig. 5** it might seem that the obvious way to build a tunnel-diode crystal oscillator is as shown in **fig. 7A.** This is

fig. 4. The Butler oscillator—a modified Colpitts for use with series-mode crystals.

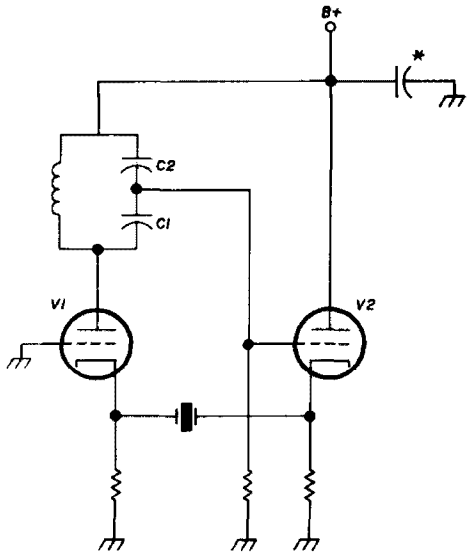


fig. 5. Negative-resistance, or two-terminal oscillator.

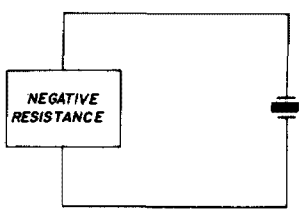
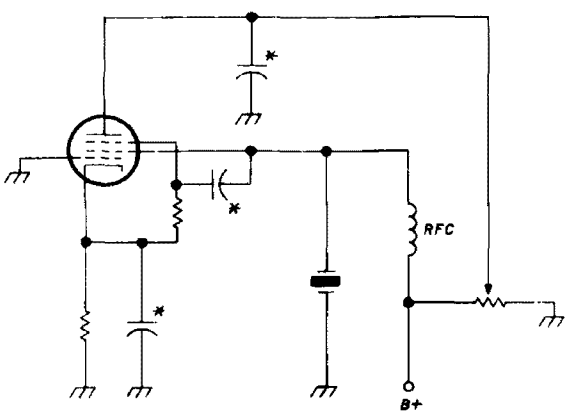
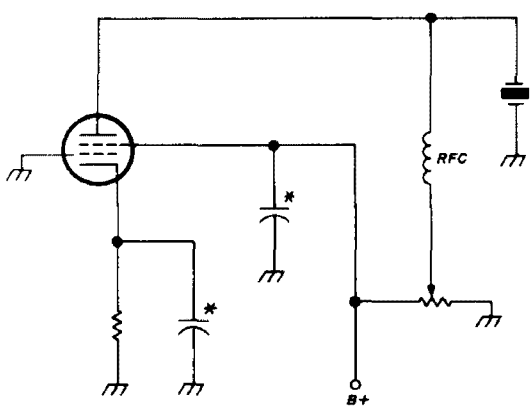


fig. 6. Transitron (A) and dynatron (B) crystal oscillators. Blocking capacitors denoted by *.



A



B

crystal unit, (military)	military holders	nominal frequency range (kHz)	frequency tolerance (± percent)	resonance	load capacitance (pF)	mode of operation
CR-1(A)/AR	HC-11 or 12/U	2,000-15,000	0.02	Parallel	35.0 ± 0.5	Fundamental
CR-2/U	FT-241A, HC-17/U	200	0.009	Parallel	125.0	Fundamental
	FT-241A, HC-17/U	500	0.010	Parallel	64.0 m	Fundamental
CR-3/U	FT-241A, HC-17/U	300-600	0.02	Parallel	—	Fundamental
CR-4/U	FT-241A, HC-17/U	500-1,200	0.02	Parallel	—	Fundamental
CR-5/U	FT-241	2,000-10,000	0.02	Parallel	25.0 ± 0.5	Fundamental
CR-6/U	FT-243	2,000-10,000	0.02	Parallel	12.0	Fundamental
CR-7/U	HC-14/U	3,750-10,000	0.004	Parallel	28.0 ± 0.5	Fundamental
CR-8/U	FT-243	1,000-10,000	0.02	Series	—	Fundamental
CR-9/U	HC-10/U	15,000-50,000	0.01	Series	—	Overtone
CR-10/U	FT-243	5,000	0.005	Parallel	25.0	Fundamental
CR-12/U	FT-243	2,000-10,000	0.02	Parallel	32.0	Fundamental
CR-13	FT-243	455	0.02	Series	—	Fundamental
CR-13/U	FT-243	5,250	0.02	Series	—	Fundamental
CR-14/U	FT-243	2,000-10,000	0.01	Parallel	32.0	Fundamental
CR-15/U	HC-5/U	80-200	0.01	Parallel	32.0 ± 0.5	Fundamental
CR-16/U	HC-5/U	80-200	0.01	Series	—	Fundamental
CR-17/U	HC-10/U	15,000-50,000	0.005	Series	—	Overtone
CR-18/U	HC-6/U	800-16,000	0.005	Parallel	32.0 ± 0.5	Fundamental
CR-19/U	HC-6/U	800-20,000	0.005	Series	—	Fundamental
CR-23/U	HC-6/U	10,000-75,000	0.005	Series	—	Overtone
CR-24/U	HC-10/U	15,000-50,000	0.005	Series	—	Overtone
CR-25/U	HC-6/U	200-500	0.01	Series	—	Fundamental
CR-26/U	HC-6/U	200-500	0.002	Series	—	Fundamental
CR-27/U	HC-6/U	800-15,000	0.002	Parallel	32.0 ± 0.5	Fundamental
CR-28/U	HC-6/U	800-20,000	0.002	Series	—	Fundamental
CR-29/U	HC-5/U	80-200	0.002	Parallel	32.0 ± 0.5	Fundamental
CR-30/U	HC-5/U	80-200	0.002	Series	—	Fundamental
CR-31/U	HC-6/U	1,000-10,000	0.005	Parallel	12.0	Fundamental
CR-32/U	HC-6/U	10,000-75,000	0.002	Series	—	Overtone

table 1. Specifications for military crystals.

similar to one shown in reference 2. While it may be possible to make such a circuit work, if just the right rf choke can be found, I had no luck with it. The circuit oscillates, but the crystal has no effect other than that of its holder capacity on the frequency. Instead of oscillating at the crystal frequency, the circuit oscillates at the series-resonant frequency of the rf choke and the diode equivalent capacity.

A more satisfactory crystal oscillator can be built as shown in **fig. 7B**.[3] At the crystal's series resonant point, the two resistors are effectively in parallel (because the crystal is a "short" between them). The two resistors must be smaller than the absolute magnitude of the tunnel-diode negative resistance yet large compared to the equivalent series resistance (at series resonance) of the crystal. Also, L and C are the same impedance as the resistors. That is, at 8775 kHz, L equals +j51 ohms and C equals −j51 ohms.

Since the equivalent series resistance at series resonance is so important to tunnel-diode crystal oscillators (and to many others as well), the resistances of different crystals for various frequencies must be known for good design. Such a listing is shown in **table 2.** Note that these figures were taken from one manufacturer's data sheets; resistances of other manufacturer's crystals will vary from these somewhat. Nevertheless, the table provides a "ball-park" figure upon which design can be started.

crystal unit, (military)	military holders	nominal frequency range (kHz)	frequency tolerance (± percent)	resonance	load capacitance (pF)	mode of operation
CR-33/U	HC-6/U	10,000-25,000	0.005	Parallel	32.0 ± 0.5	Overtone
CR-35/U	HC-6/U	800-20,000	0.002	Series	—	Fundamental
CR-36/U	HC-6/U	800-15,000	0.002	Parallel	32.0 ± 0.5	Fundamental
CR-37/U	HC-13/U	90-250	0.02	Parallel	20.0 ± 0.5	Fundamental
CR-38/U	HC-13/U	16-100	0.012	Parallel	20.0 ± 0.5	Fundamental
CR-39/U	HC-15/U	160-330	0.004	Series	—	Fundamental
CR-40/U	HC-15/U	160-330	0.003	Series	—	Fundamental
CR-42/U	HC-13/U	90-250	0.003	Parallel	32.0 ± 0.5	Fundamental
CR-43/U	HC-16/U	80-860	0.035	Parallel	45.0 ± 1.0	Fundamental
CR-44/U	HC-6/U	15,000-20,000	0.002	Parallel	32.0 ± 0.5	Fundamental
CR-45/U	HC-6/U	455	0.02	Series	—	Fundamental
CR-46/U	HC-6/U	200-500	0.01	Parallel	20.0 ± 0.5	Fundamental
CR-47/U	HC-6/U	200-500	0.002	Parallel	20.0 ± 0.5	Fundamental
CR-48/U	HC-6/U	800-3,000	0.0075	Parallel	32.0 ± 0.5	Fundamental
CR-49/U	HC-6/U	800-3,000	0.0075	Parallel	32.0 ± 0.5	Fundamental
CR-50/U	HC-13/U	16-100	0.012	Series	—	Fundamental
CR-51/U	HC-6/U	10,000-61,000	0.005	Series	—	Overtone (Pressure)
CR-52/U	HC-6/U	10,000-61,000	0.005	Series	—	Overtone (Plated)
CR-53/U	HC-6/U	50,000-87,000	0.005	Series	—	Overtone (Pressure)
CR-54/U	HC-6/U	50,000-87,000	0.005	Series	—	Overtone (Plated)
CR-55/U	HC-18/U	17,000-61,000	0.005	Series	—	Overtone
CR-56/U	HC-18/U	50,000-87,000	0.005	Series	—	Overtone
CR-57/U	HC-6/U	500	0.001	Parallel	32.0 ± 0.5	Fundamental
CR-58/U	HC-17/U	3,000-20,000	0.005	Parallel	32.0 ± 0.5	Fundamental
CR-59/U	HC-18/U	50,000-91,000	0.002	Series	—	Overtone
CR-60/U	HC-18/U	7,000-20,000	0.005	Series	—	Overtone
CR-61/U	HC-18/U	17,000-61,000	0.002	Series	—	Overtone

the multivibrator

The multivibrator crystal oscillator didn't really come into its own until the advent of bipolar transistors, although it has been built using tubes. It is simply an astable multivibrator with one of its coupling capacitors replaced by a crystal (or with a crystal and series-tuning capacitor). **Fig. 8** is representative of such an oscillator. This circuit was used as a one-megahertz crystal calibrator in vhf receivers.[4]

The multivibrator crystal oscillator isn't particularly noteworthy as shown (built with discrete components), but it's the basis for most of the crystal oscillators using digital integrated circuits.

The main use of multivibrator-type crystal oscillators has been for frequencies be-

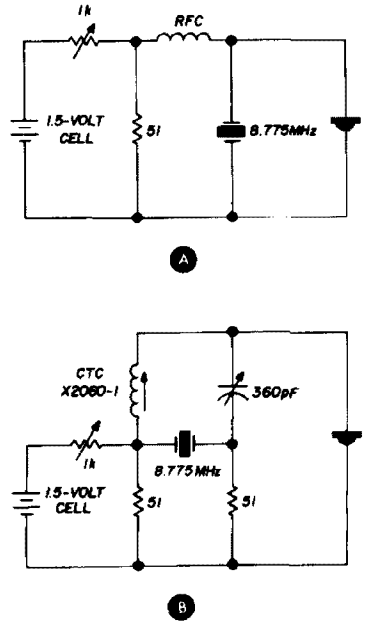


fig. 7. Tunnel-diode oscillators. Circuit at (A) is not recommended because of critical adjustment of rfc.

table 2. Series-resonant resistances of crystals for different frequencies.

Reevas-Hoffman type	Frequency (kHz)	Series-resonant Resistance (ohms)	Remarks
J element	1	150,000	+5° × cut
J element	2	75,000	+5° × cut
J element	4	36,000	+5° × cut
J element	8	30,000	+5° × cut
J element	12	15,000	+5° × cut
N element	20	9,000	NT cut
N element	50	10,000	NT cut
E element	100	1,800	+5° × cut
D element	200	950	Y cut
C element	500	1,000	Y cut
A element	1×10^3	300	AT cut (fundamental)
A element	3×10^3	65	AT cut (fundamental)
A element	6.5×10^3	15	AT cut (fundamental)
A element	1.15×10^4	12	AT cut (fundamental)
A element	2.08×10^4	8	AT cut (3rd overtone)
A element	4.04×10^4	35	AT cut (3rd overtone)

low 100 kHz, where it's impractical to provide large inductors for conventional oscillators.

using bipolar transistors

The translation of tube-type crystal oscillators to solid-state circuits must be approached with some caution. Too often a bipolar transistor symbol is just drawn in place of a triode tube symbol. This sort of "engineering" has resulted in circuits like that of **fig. 9**, which **sometimes** oscillate.

A bipolar transistor is quite unlike a vacuum tube; a **current** into the base controls the collector current in the transistor. In a tube, a **voltage** on the grid controls the plate current. As a result of these different behaviors, the base of a common-emitter bipolar transistor has a much lower input impedance than the grid of a grounded-cathode vacuum tube.

The lower impedances presented by the bipolar transistor make it difficult to use in crystal oscillators that use parallel-resonant crystals. However, for use with series-resonant crystals, the bipolar transistor is just great. **Fig. 10** shows how series-mode crystals and bipolar transistors can be used to take advantage of the compatible impedances of each.

Note also in the circuits of **fig. 10** that a 6.8-μH inductor is placed in parallel with the 26-MHz crystal. This inductor forms a parallel-resonant circuit with the holder capacitance of the crystal. Looking at **table 1**, the CR24/U has a holder capacitance of 7 pF. At 26 MHz, approximately 6 μH resonates 7 pF, so the closest

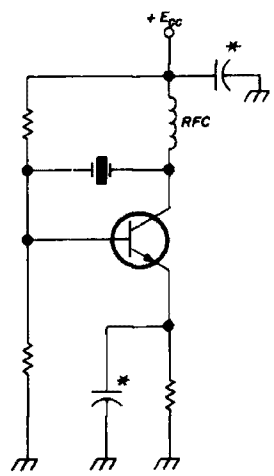
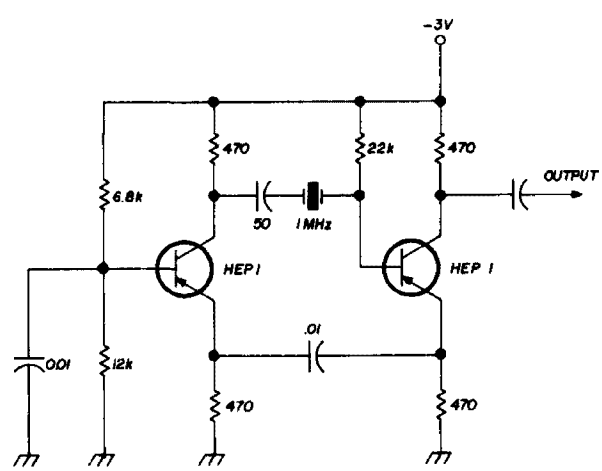


fig. 9. Pierce oscillator using a bipolar transistor—not recommended (see text). Blocking capacitor denoted by *.

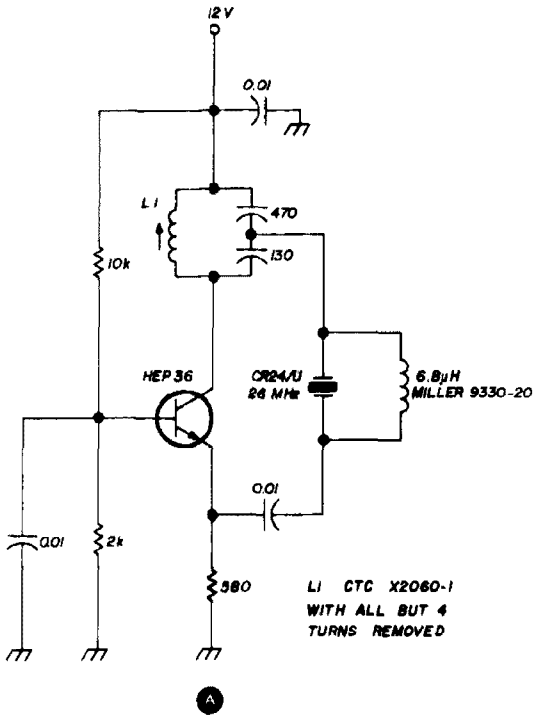
fig. 8. Multivibrator crystal oscillator. The HEP 1's can be replaced with 2N1204's.



standard inductor (6.8 μH) in the Miller 9330 series was used.

Balancing the holder capacitance in this way acts as a simple form of mode filter, assuring that the crystal oscillates as marked. I found this simple method to be adequate even with seventh-mode overtone crystals in the 170-MHz range. The method can be applied to nearly any crystal oscillator where the crystal is to be operated in the series mode. However, the addition of a series-blocking capacitor may be necessary (since the **crystal** may have blocked dc in the original circuit).

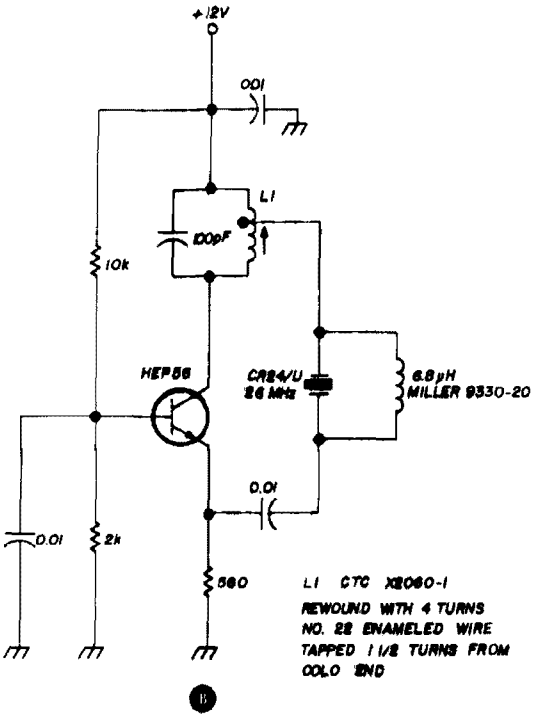
fig. 10. Series-mode crystal oscillators using a bipolar transistor—Colpitts (A), Hartley (B) and tickler-feedback (C).

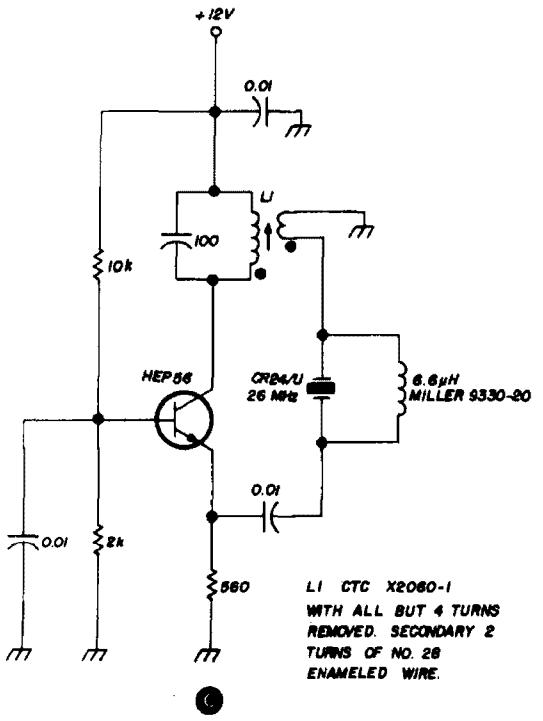


fet crystal oscillators

These devices, on the other hand, are quite similar to tubes because the gate voltage controls drain current in the common-source connection. So almost any vacuum-tube crystal oscillator finds its direct equivalent in an fet circuit. Two are illustrated in **fig. 11.** Notice that both use junction fet's. This was done because the diode junction between gate and source conducts (like the grid and cathode of a tube) when forward-biased, and allows "gate leak" action. If an insulated-gate field-effect transistor (igfet) is used, a separate diode is usually added across gate and source, as in **fig. 12.**

Fig. 12 is an ultra-simple Pierce oscillator (a version of the Colpitts), which finds





its counterpart vacuum-tube circuit in **fig. 2A**. It's useful for checking the general activity of fundamental-mode crystals in the 2- to 20-MHz range. Since fundamental-mode crystals aren't usually cut above 20 MHz, if a higher frequency rock is plugged in (say a 26-MHz, 3rd-mode) it won't oscillate as marked. One indication of oscillation in this circuit is a decrease in drain current, since the nonoscillating circuit draws I_{DSS} until gate-leak action is established.

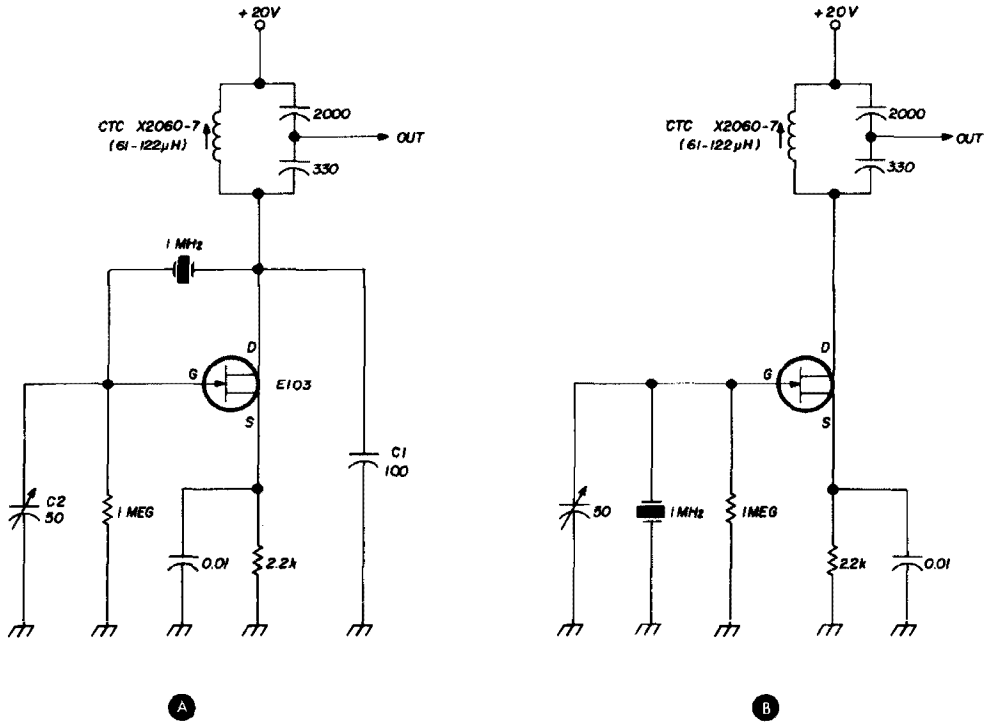


fig. 11. Crystal oscillators using n-channel junction fet's. In (A) crystal performs as an inductor between drain and gate of the fet, with capacitive divider formed by C and C2. Circuit (B) is the Miller oscillator, which is useful with crystal having one terminal connected to the can (as in DC-9 octal style).

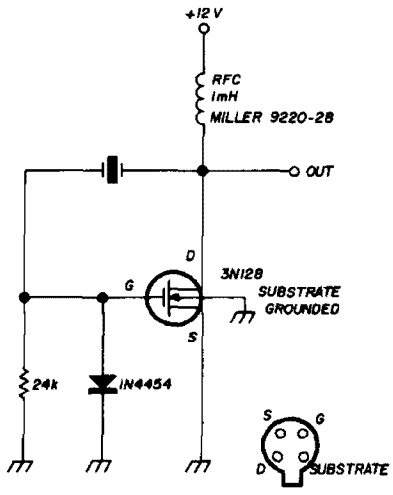


fig. 12. Pierce circuit using n-channel igfet; useful with fundamental-mode crystals from 2 to 20 MHz.

No circuits using p-channel fet's are shown (either junction or insulated-gate types) because rf types haven't become available at the low prices of the n-channel units. The germanium TIXM12 was one of the exceptions (a p-channel JFET, good to over 100 MHz for $1.07), but it has been discontinued like the TIXM05 we all loved. If and when p-channel fet's for rf become available again, the circuits of **fig. 11** and **12** can be used by reversing the supply-voltage polarity (and reversing the diode in the circuit of **fig. 12**).

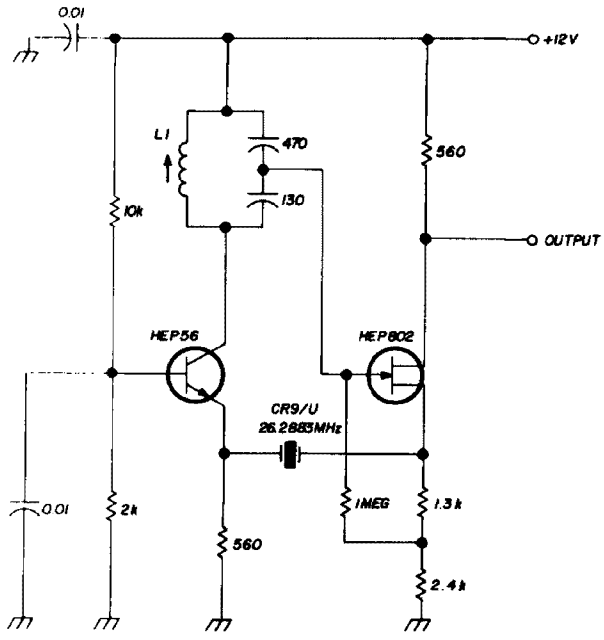
There are, of course, a number of ways in which npn and pnp bipolar transistors can be combined with n-channel and p-channel jfet's and igfet's to provide com-

bination circuits. An example is shown in **fig. 13**, a Butler oscillator using an npn bipolar transistor and an n-channel jfet.

integrated circuits

In recent years, one of the largest areas of growth in the semiconductor industry has been in integrated circuits. These can be used as crystal oscillators in a number of configurations. Digital IC's, which are the least expensive units, can be made to function as crystal oscillators of the multivibrator-type. **Figs. 14** through **17** show the RTL, DTL, TTL, and ECL families as crystal oscillators. Resistor-transistor logic (RTL) is the least expensive family and has been widely used in amateur systems because of its low cost. It is relatively slow, however, and the circuit of **fig. 14** can't be expected to work reliably at frequencies higher than a few megahertz.[6] By using a pair of the higher-power μL900 buffers, oscillation can be obtained up to about 8 MHz.[7] The circuit of **fig. 15** apparently operates at higher frequencies because of lower inherent resistances.

Although RTL is the least expensive, the prices of diode-transistor logic (DTL) and transistor-transistor logic (TTL) have been steadily decreasing, and they're now feasible for ham construction. A cost comparison of several quad-dual gates shows the price per gate to be within reason (**table 3**).



L1 = CTC X2060-1 WITH ALL BUT 4 TURNS REMOVED

fig. 13. Butler oscillator using npn transistor and n-channel fat.

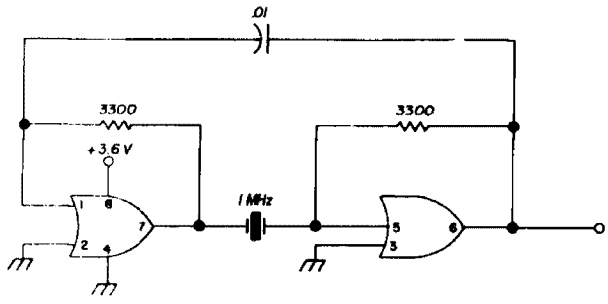


fig. 14. Crystal oscillator using resistor-transistor logic (RTL) integrated circuit.

fig. 15. Using RTL buffers in a crystal oscillator circuit.

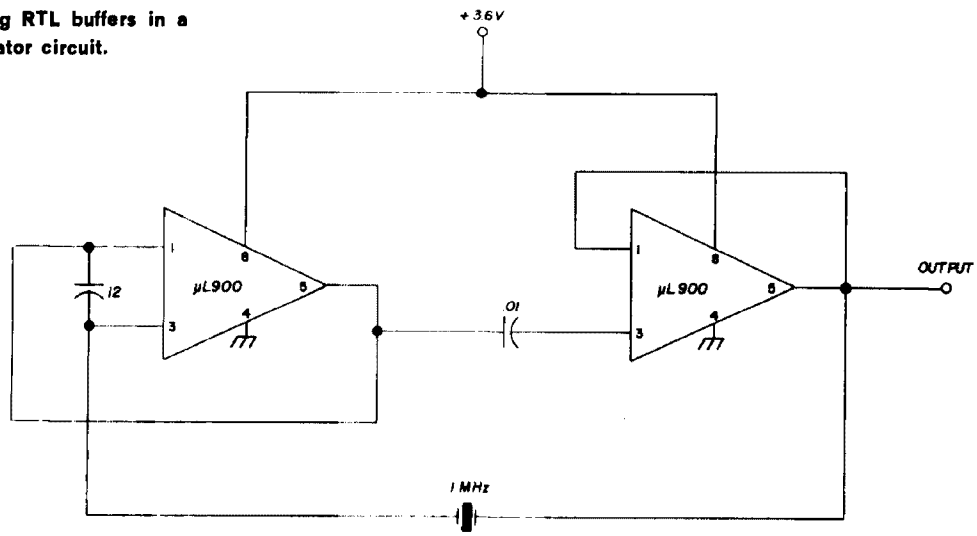
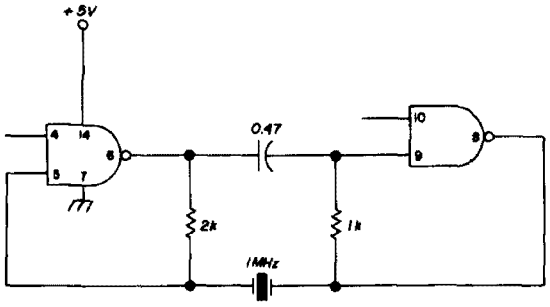


tabla 3. Cost comparison of popular integrated circuits.

Logic	Company	Part Number	Cost	Cost Per Gate
RTL	Motorola	MC717P	$1.08	$0.27
DTL	Motorola	MC846P	1.65	.41¼
TTL	TI	SN7400N	2.25	.56¼
ECL	Motorola	MC1010P	1.80	.45

Fig. 16 shows how DTL and TTL units can be used as a crystal oscillator. Note that the same pin configuration and voltage are used for the two IC's.[6] Also, since only two of the four gates are used, the others can be used as isolation stages or as another crystal oscillator.

One word of caution about use of TTL's, however: the pins shown are **only** for the SN7400N. If you try the military version (SN7400 or SN5400), pin connections are quite different!



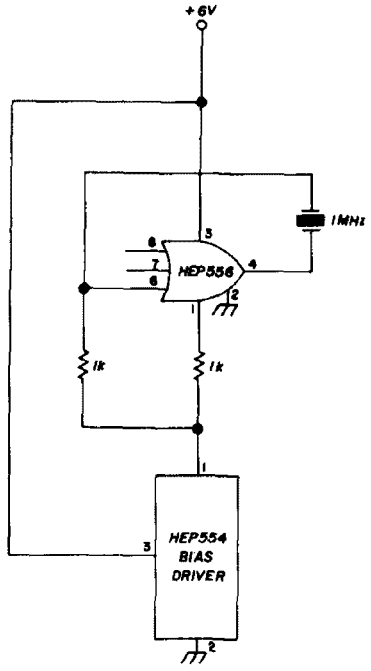
IC IS SN7400N (TEXAS INSTRUMENTS)
OR MC846P (MOTOROLA)

fig. 16. Transistor-transistor logic (TTL) or diode-transistor logic (DTL) gates in a crystal oscillator.

The emitter-coupled logic (ECL) family is the only one where the internal transistors aren't switched into and out of saturation. This feature makes ECL inherently fast, allowing the MECL I series of Motorola to operate to 30 MHz, and the MECL II series to approach 100 MHz as crystal oscillators. **Fig. 17** shows how an MECL I gate can be used as a simple 1-MHz oscillator.[8] Reference 9 covers in more detail the ECL at higher frequencies.

A separate bias driver (MC354G or HEP554) is required with the MECL I series, but it's built in on the MECL II series. The cost per gate of MECL II, using the MC1010P (quad 2-input gate, at $1.80) is $0.45. This compares closely with DTL and TTL units.

fig. 17. Emitter-coupled logic (ECL) gates used in a crystal oscillator.



the operational amplifier

In linear IC's there is almost limitless variety, but one of the main building blocks is the direct-coupled differential amplifier. Two forms of this monolithic IC have become more or less standard in the semiconductor industry: the operational amplifier (like the Fairchild μA709 and its descendants) and the rf/i-f amplifier (the Fairchild μA703 and similar units). Both can be used as crystal oscillators; the operational amplifier is limited to the lower frequencies.

An operational amplifier used as an oscillator brings out many of the basic fundamentals of oscillator design. Because the operational amplifier is such a nearly ideal device, it affords ease of feedback design. Inverting and noninverting inputs are provided on most op-amps, both positive feedback (used to cause oscillation)

and negative feedback (used to reduce gain and stabilize output) can be selectively used to produce exactly the oscillator design you desire.

An example of an op-amp crystal oscillator using both positive and negative feedback is shown in **fig. 18**. This circuit uses a nonlinear resistor in the negative feedback resistive network to adjust the gain and thereby assure sinusoidal waveform. The technique is similar to that used in the Wien Bridge audio oscillator, and is covered in reference [10].

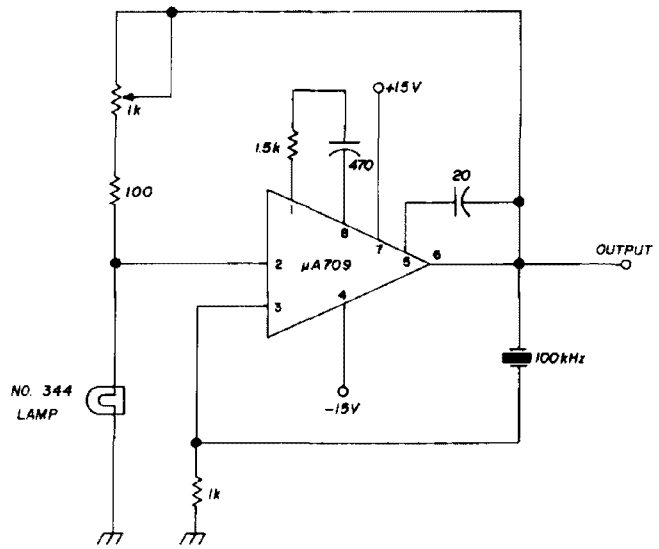


fig. 18. Monolithic operational amplifier as a crystal oscillator.

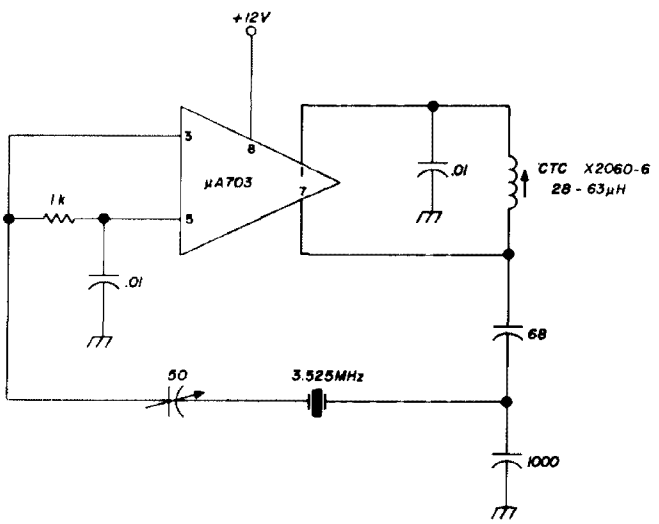


fig. 19. The μA703 as a 3.525-MHz crystal oscillator.

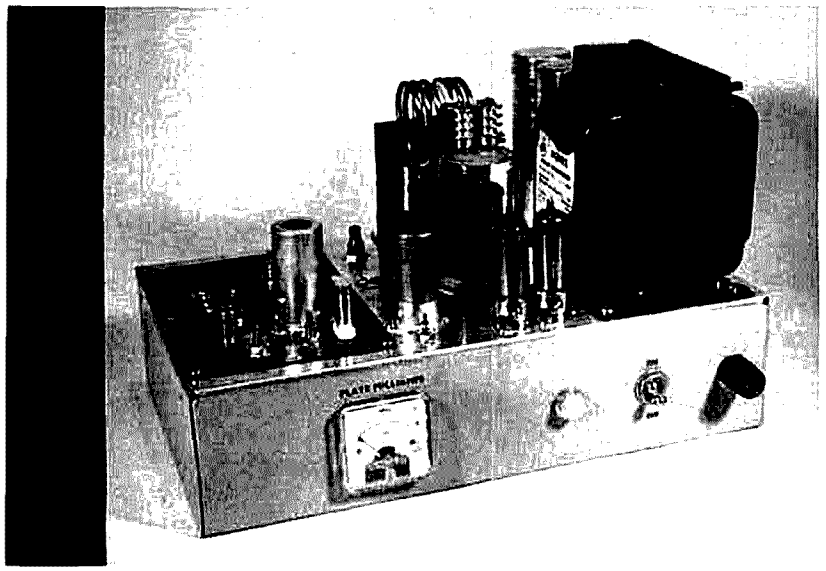
A number of other linear integrated circuits have been used as crystal oscillators and are described in various application notes.[11,12,13] The main impetus for the use of IC's in crystal oscillators has been in 3.58-MHz TV colorburst generators; so most of the circuits shown in these references are for 3.58-MHz oscillators.

You might ask what advantage these IC crystal oscillators have over those built from discrete components—if any. The answer is that there's little advantage in using IC's in this way, unless you consider it avante-garde to have **your** piece of equipment "all IC." The IC **does** offer a rather large stable gain in one package, however, as evidenced by the large capacitive divider across the tuning coil.

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Every time the six-meter skip comes in you're no doubt impressed by the rapidly increasing number of single sideband stations. "Maybe it's time to plan for that sideband rig," you think. Then you look at the price tags on the commercial outfits and decide, "Not yet." Even the kits are expensive, and whipping one up from scratch is both expensive and complex.

A much more attractive answer to a good sideband signal on six meters is a combined receiving and transmitting converter. If you have a low-band ssb transceiver, a grid-dip meter and a little homebrew experience, this can be a very happy solution to the problem.

I went through these mental gymnastics for a couple of years before deciding to take soldering gun in hand and build the **6/40 'verter.** One of Mr. Heath's excellent Single-Banders for forty meters was in the shack. I put together a conventional converter to shift the six-meter sideband signal down to forty. You've probably had some experience along this line and this is half the job. A somewhat similar converter was built alongside to swing the forty-meter ssb signal from the transceiver up to six meters. A simple

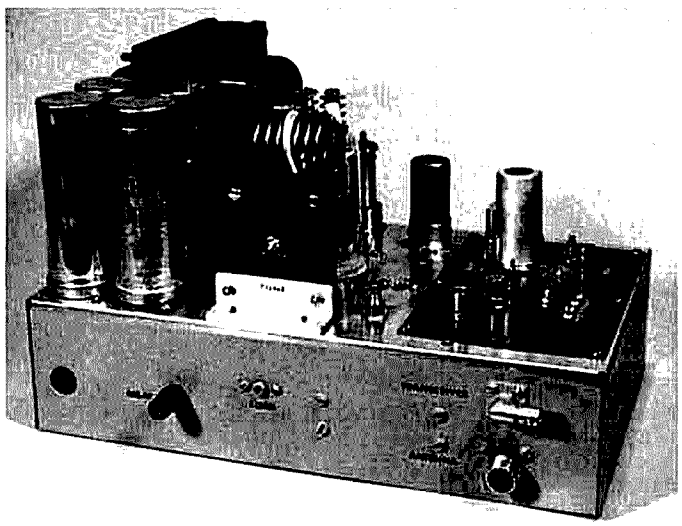
power supply for both converters was added, and the job was done. Both converters use the same oscillator, so both the received and transmitted signals are on the same frequency. Once tuned up, the transverter can be disregarded, and operation is with the transceiver only.

Right about here you're probably saying, "It can't be that simple. The generation of ssb signals is complex, and homebrewing of such equipment just isn't for me." You're absolutely right. But, the difficult part has all been taken care of by those smart people at Benton Harbor. All we're talking about here is beating signals to produce heterodynes, as in that ten-dollar table radio in the bedroom. The fact that one of the signals is modulated ssb instead of a-m doesn't change the operation.

the circuit

With this description in mind let's examine the schematic (**fig. 1**). Note that the top half, consisting of the 6CW4, 6EA8 and associated circuit is the familiar receiving converter. The one used here isn't original—it's borrowed from "The Radio Handbook." It's a good one, but there's no reason to use this if you have a converter on hand or wish to buy a commercial unit.

For transmitting, the conversion signal from the triode section of the 6EA8 is fed to a 6DJ8 cascode amplifier. The boosted rf is then inserted push-pull into the grids of a 5894 serving as a mixer-final. The forty-meter ssb signal from the Single-Bander is inserted at the screen grids of the 5894, and a six-meter, single-banded signal comes out. The 5894 dual tetrode is a relatively high-power, easily driven tube that isn't exactly cheap, but it can be found in some of the bargain-priced tube lists.



It does a beautiful job in this application, but maybe you have an 815 or 6360, or a couple of 2E26's you'd like to use. A few component values would require changing, but these tubes would all work—at lower power. By the same token, there's no reason why you can't start with a twenty-meter signal and end up on two meters. The approach would be the same. It all depends on what you have to build with and where you want to operate.

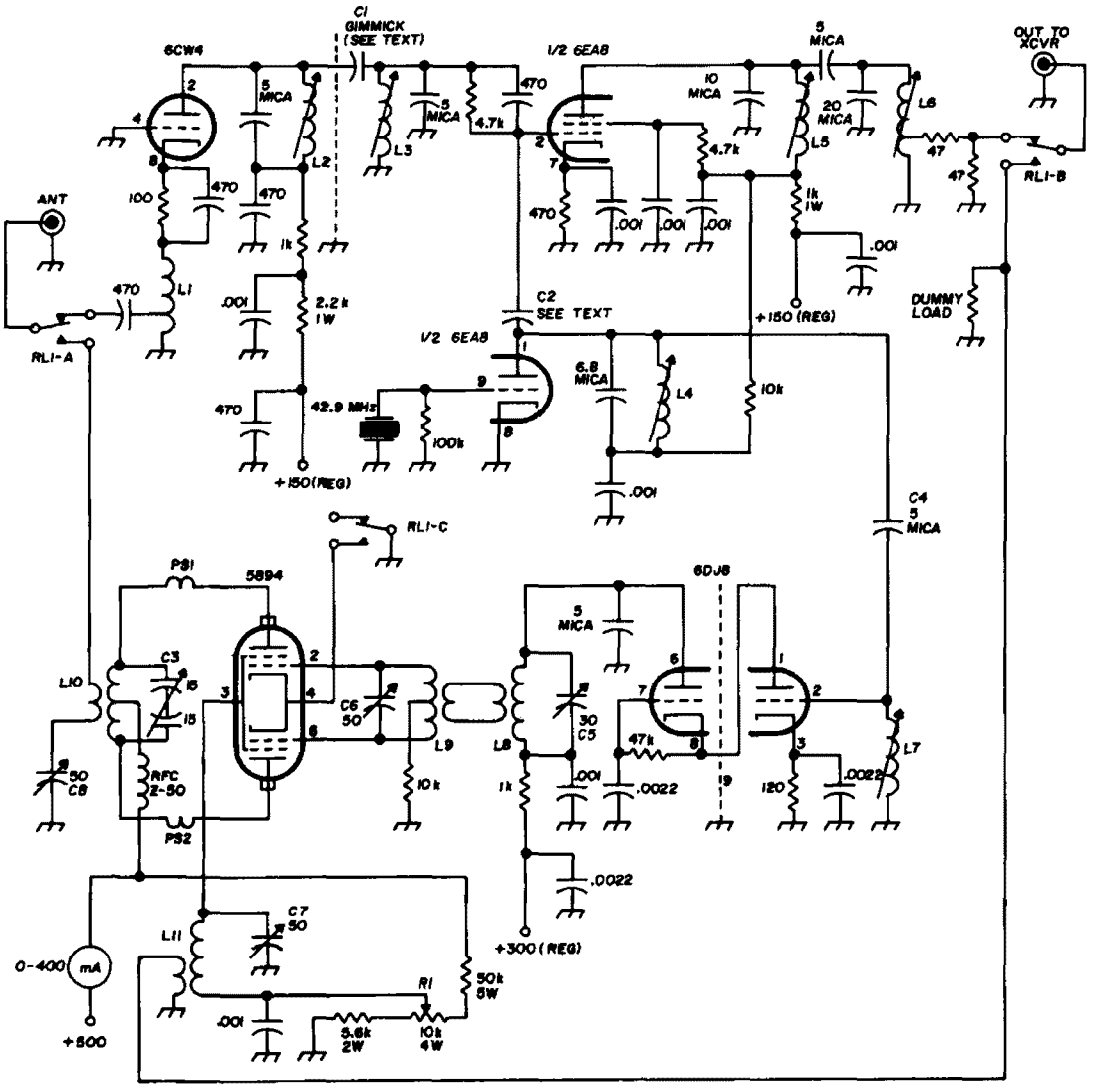
assembly

Construction practices are standard for these frequencies. Good grounds, short leads, many bypasses and shielding are very important. The layout shown in the photos is compact, and it works, but it isn't necessarily the last word. The receiving converter was built on a four- by six-inch sheet of flashing copper. Good rf grounds are no problem since many of the components can be soldered directly to the copper.

The 6CW4 is a grounded-grid nuvistor rf amplifier. It's lightly coupled to the 6EA8 mixer section via C1, which consists of two lengths of hook-up wire with about one inch of each parallel with the other, inserted from opposite sides through a hole in the shield between the two tubes. Coupling between oscillator and mixer portions of the 6EA8 is via the spacing of about 3/4 inch between L3 and L4. The coax input and output fittings can be eliminated—they're a reminder of when this converter was used alone.

The same crystal oscillator signal used for receiving is taken from the plate of the converter portion of the 6EA8 via C4 to the first grid of the 6DJ8. Here it goes through two stages of amplification needed for the transmit conversion. The 6DJ8 output is inductively coupled to the 5894 signal grids.

A very low level forty-meter signal is required. This is achieved by dissipating most of the output in a dummy load. A sur-

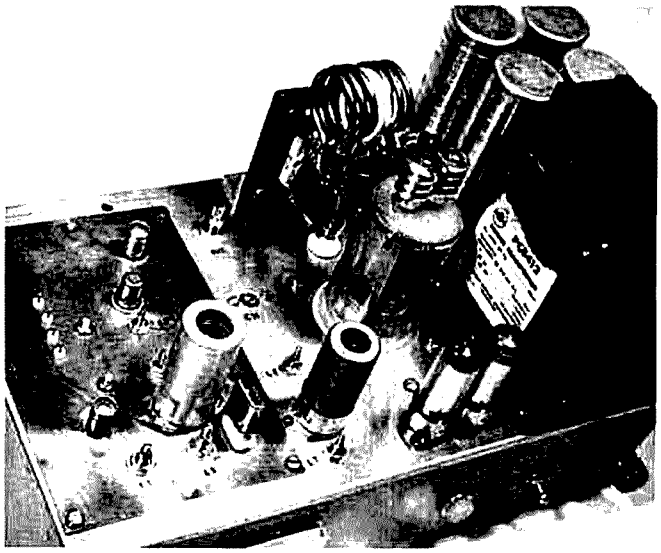


C1	0.5 pF gimmick (see text)
C2	capacitive pickup between L3 and L4
C3	15 pF butterfly
L1	10 turns B&W 3003, tapped 2½ turns from ground end
L2, L3	0.87 μH (J. W. Miller 40A827CBI)
L4	1.0 μH (J. W. Miller 40A106CBI)
L5, L6	Miller 41A000CBI wound full with no. 26 enamelled; L16 tapped 15 turns from ground end
L7	2.2 μH (J. W. Miller 41A226CBI)
L8	8 turns no. 28 enamelled on a ¼" slug-tuned form (J. W. Miller 41A000CBI); link is 2 turns of hookup wire on the cold end
L9	9 turns B&W 3004, center tapped with 2 turns of hookup wire around center
L10	6 turns no. 12, 1-inch ID, spaced 1 wire diameter, with 2-turn insulated loop at wider-spaced center
L11	48 turns B&W 3004; link is 4 turns of hook-up wire around cold end
PS1, PS2	6 turns no. 18, spaced one wire diameter, on high-value, 2 W resistors

fig. 1. Schematic diagram of the 6/40 'verter.

plus three-pole, double throw dc relay switches the antenna circuit alternately through the receiving and transmitting converter sections and turns on the final for transmit by grounding its cathode. Any similar commercial relay will do this job, and by choosing a six-volt ac coil, the rectifier diodes and filter capacitors, which can be seen in one of the photos, can be eliminated. The Single-Bander has an extra set of relay contacts that activate the transverter relay to make the switching automatic.

The entire outfit, including power supply, fits neatly on a commercial 3x7x12-inch chassis base. A hole was cut in the top at one end to accept the copper sheet of the receiving converter. The photos show the placement of the other components. Note that C3 is ungrounded. It mounts on a vertical square of phenolic board to which the output winding of L10 is also attached. The parasitic chokes for the 5894 are soldered across loops in the copper straps connecting its plates to C3 and L10.



Top view showing the general layout with the receiving converter at the left on its copper plate, the transmitting converter in the center, and power supply on the right.

The power supply is conventional (fig. 2) It provides 500 Vdc at 300-plus mA for the 5894 (which, incidentally, can be pushed much harder); 300-volts regulated for the 6DJ8; and 150-volts regulated for the 6CW4 and 6EA8. Signal stability is essential on single sideband. This depends on your transceiver, but the voltage regulation provided here ensures against any frequency drift in the transverter. Capacitors C1 and C3 are mounted on insulated bases, and the cans are insulated to prevent accidental shock.

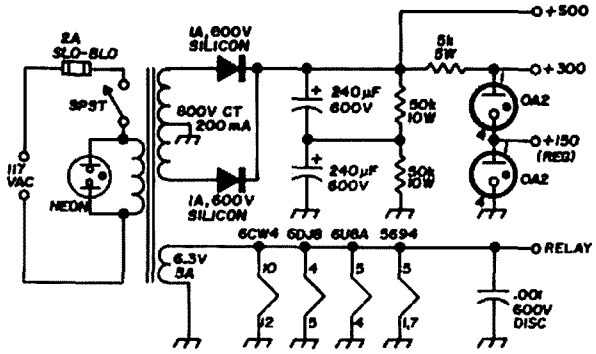
appearance

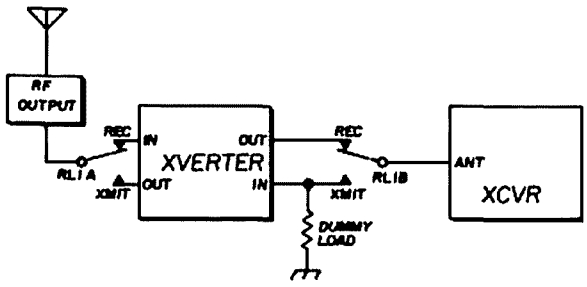
Finishing touches include a ventilated top cover, formed from sheet aluminum, and a ventilated bottom plate. The shielding is required for best results. A coat of spray paint and decal identification of controls will give a professional appearance.

tuning up

The forty-meter Single-Bander tunes only 100 kHz—from 7.2 MHz to 7.3 MHz. Thus, a 42.9-MHz crystal will cover 50.1 to 50.2 MHz. This is the portion of six meters where

fig. 2. Power supply for the transverter and relay switching arrangement. Relay is a 6.3 Vac unit.





Relay switching arrangement.

most ssb signals are heard. Crystals can be chosen, however, to put you anywhere you want in the band.

Assuming you plan to operate 50.1-50.2 MHz, a grid-dip meter can be used to resonate L2 and L3 a little under, and a little over, 50.15 MHz. L4 is tuned to the crystal frequency; 7.25 MHz is bracketed with L5 and L6. This should give a good, flat response across the 100 kHz used. When you get far enough along to copy signals over the air, these adjustments can be touched up to produce the best signal-to-noise ratio.

L7, C5-L8, and C6-L9 are resonated to the crystal frequency. C7 is adjusted to peak L11 at 7.25 MHz. Adjust C3 and the spacing of L10 to 50.15 MHz. By making these adjustments to the center of the 100-kHz segment of interest, it will be unnecessary to do any further tuning of the transverter as you change frequency.

on-the-air tests

After checking and rechecking your work, you're ready for the smoke test. Connect your **6/40 'verter** between your transceiver and your six-meter antenna. Turn on the power and listen for signals. Listen on the upper sideband, because stations on six are almost always on upper sideband and those on forty are on lower sideband.

With the receiving converter working, adjust L4 for maximum using your dipper as an absorption wavemeter. Move to L8 and adjust L7 and C5 for a peak, which will be very much stronger. Touch up C6 for a high reading at L9.

Now, with no forty-meter input, close the relay manually and adjust R1 for a plate current of 50 to 75 mA. Connect a dummy load in the **transceiver** output to reduce the forty-meter signal to one or two watts, and attach a dummy load to the **transverter** output. Insert a power-output meter between transverter and dummy load. Interconnect the relays so the transceiver and transverter transmit together. Turn the transceiver to **tune** position to provide a carrier, and adjust C3 and C8 for maximum output. C7 and R1 can be readjusted to peak the output. During tune-up, the injected carrier level should be low enough to keep the 5894 plate current under 200 mA. **Don't let this tube run red.**

Return the transceiver **tune** control to transmit, and speak into the mike. The plate meter should kick on voice peaks to approximately 300 mA. This completes tune up, and you're ready for on-the-air tests. Listening tests will provide the best criteria of correct adjustment. Important: correct adjustment should occur when the **least** amount of ssb drive, as established with R1 and C7, produces the greatest output. The quality of your six-meter sideband signal should be as good as your forty-meter signal. Reports for the **6/40 'verter** have been uniformly good.



The ventilated cover improves appearance and provides essential shielding and protection from accidental contact with the high voltage.

a final note

This probably shouldn't be your first construction project, but you don't have to be an engineer to achieve good results. Single sideband isn't all that difficult when you do it the easy way. The idea is to proceed slowly, plan each step, and follow my instructions and suggestions. If you can solder and use simple shop tools, you can build the **6/40 'verter** in a few week ends.

ham radio

stub bandswitched antennas

Two multiband verticals are described: a fixed-station antenna and a twinlead portable— no loading coils; no traps

John J. Schultz, W2EEY, 40 Rossie Street, Mystic, Connecticut 06355

One of the problems in designing vertical antennas is finding a simple bandswitching method. To decouple portions of the antenna for different bands, you can use (a) basemounted switched loading inductors, or (b) traps. Each method has disadvantages. A remotely controlled switching system adds extra wiring and cost, and traps are difficult to build, adjust and mount.

This article describes a simple and inexpensive method of bandswitching using the *principle* of stub decoupling. I used it with two antennas: a portable multiband wire antenna and a fixed-installation vertical. The idea can be extended to many other antennas as well.

stub switching

A 33-foot vertical is shown in **fig. 1A.** This antenna would function as a quarter-wave vertical on 40 meters. If a stub approximately 15½ feet long were added, as shown in **fig. 1B**, tri-band operation would be possible.

Operation on the various bands is as follows. On 40 meters, the stub is too short to have any effect, and the antenna performs as a simple ¼-wavelength vertical. On 20 meters, the 15½-foot stub is a quarter wavelength long; the short circuit at the base of the stub is reflected as an open circuit at the high end of the stub. Thus the upper portion of the 33-foot vertical is decoupled, and the antenna performs as a ¼-wavelength vertical on 20 meters.

On 15 meters, the 15½-foot stub has no switching effect since it is neither ¼-wave-

length or 1/2-wavelength long. The 33-foot section is active, and the antenna functions as a 3/4-wavelength vertical. Such a length will present a low impedance at the antenna base to match a coaxial transmission line.

There is some disadvantage to this antenna length, however. It is slightly longer than the optimum length for maximum low-angle radiation. Therefore, some high-angle radition will also occur on 15 meters.

Still another stub can be used to extend operation to 10 meters. This stub can be placed in a number of ways, but the most advantageous arrangement seems to be that shown in **fig. 1C.** An approximate 8-foot stub is placed about 8 feet from the base of the antenna. The stub acts as a phase reversal device to couple the lower 8-foot section (1/4 wavelength on 10 meters) to the upper 16-foot section (1/2 wavelength on 10 meters) on the main antenna. A collinear vertical array results as a consequence of the phase reversal. This keeps the antenna radiation at a low vertical angle and produces a slight gain (1 to 2 dB) over a simple 1/4-wavelength vertical. The 15 1/2-foot stub has no effect on 10 meters since it is 1/2-wavelength long and simply reflects a short circuit at its upper end. The 10-meter stub has no effect on operation on other bands.

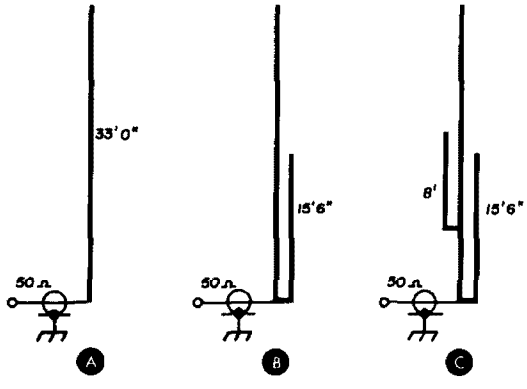


fig. 1. Various arrangements with a 33-foot vertical antenna. Simple vertical in A performs on 40 and 15 meters. The 15 1/2-foot stub in B adds 20-meter capability. Another stub, approximately 8 feet long, makes the vertical operational on all bands from 40 through 10 meters.

The over-all result is a 4-band antenna that performs as a 1/4-wavelength vertical on 2 bands, a 3/4-wavelength vertical on 1 band, and a collinear array on the highest-frequency band. The dimensions of the antenna differ slightly from those of a basic 1/4-wavelength antenna, because the stubs affect the diameter-to-length ratio of the antenna on the various bands. Provision must be made during construction for adjusting the antenna element lengths and for initial stub placement in some cases.

fixed vertical antenna construction

I experimented first with a tri-band antenna of the type shown in **fig. 1B.** A multisection telescoping aluminum element was used for the antenna section, which had a maximum diameter of 1 1/4 inches. The stub was made of similar material but with fewer sections. Spacing between stub and antenna was determined by the type of insulating spacer used as shown in the photograph; about 4 1/2 inches between stub and main-element centers.

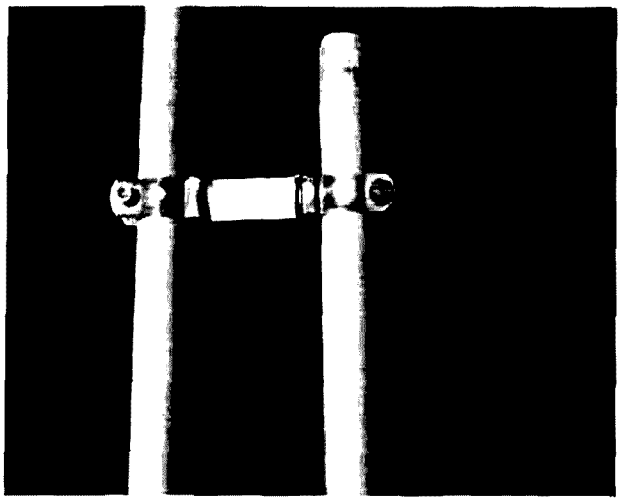
Birnbach Company produces a series of pillar insulators that can be used for stub holders of almost any desired size. Their type 445H, for instance, is 3 inches long and 3/4-inch in diameter. Both ends are threaded for 10-32 hardware. The holders on each end of the pillar insulator can be purchased or you can make your own.

At the base of the antenna, the stub and main element ends are joined by a piece of Belden braid. The center conductor of the 50-ohm coax is connected to this braid and the shield of the coaxial cable to another braid, which connects two 6-foot ground rods spaced 3 feet apart, centered on the antenna base. I used only ground rods because the soil is fairly moist in the vicinity of the antenna. Dry locations will require a radial system. The antenna is physically supported by a wooden post. There is nothing special about the construction, and almost any method for vertical antenna construction can be used.

Little adjustment is required for tri-band operation. With the antenna excited on 40 meters, the main element length is adjusted for the lowest swr in the transmission line. A value of 1.5-to-1 or less should result.

Switching to 20 meters, the stub is adjusted for the lowest swr. No significant interaction should occur between these adjustments unless you started with element lengths that were out of resonance. Once adjusted on 40 meters, the antenna should be properly tuned on 15 meters to a compromise setting for coverage between 40- and 15-meter band segments.

Pillar insulators are used between the stubs and the main antenna element.



If an additional stub is added for 10-meter coverage, the preceding adjustments should be made before the stub is mounted. Then the stub should be placed about 8 feet from the base. With the antenna excited on 10 meters, both stub location and length should be varied slightly until the lowest swr is achieved. It would be useful to have another station or a field-strength meter a few wavelengths away to help indicate the stub location for maximum signal.

Finally, the antenna should be rechecked on the other bands to ensure that no significant detuning has occurred. If so, a back-and-forth tuning procedure must be used until proper tuneup is achieved on all bands.

twinlead stub antenna

The basic simplicity of the multi-band vertical stub system produced the idea for a similar antenna made only of 300-ohm twinlead. The twinlead antenna was made for portable use as a multiband antenna that could be strung up and used without an antenna coupler.

Since the twinlead provides only two conductors, a somewhat different stub was used (fig. 2). The conductors are connected together at the far end of the antenna, and one conductor is cut a little less than 16 feet from the far end to form the 20-meter stub. This stub is 1/4-wavelength long on 20 meters and reflects an open circuit, so the lower portion acts as a 1/4-wavelength antenna on 20 meters.

On 40 meters, the 20-meter stub adds some top loading, but essentially it's not active, and the unbroken conductor forms a 1/4-wavelength antenna. A 20-meter stub can be added in the same manner as for the fixed-station vertical by using the remaining twin-lead conductor after forming the 20-meter stub. On 15 meters the 40-meter section is used as a 3/4-wavelength antenna.

tuning up

Because of the twinlead velocity factor and the nature of the stub arrangement, pruning this antenna is a bit more tedious than the fixed-station vertical. However, twinlead is inexpensive, and even if you foul up the tuning the first time around no great loss will result. In fact, experimenting with the twinlead version is a good way to gain confidence in the basic antenna operation before constructing a more expensive fixed-station version.

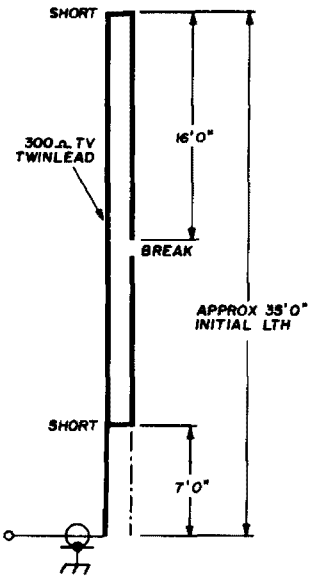
As a first step in the tuning process, choose a 35-foot length of twinlead and leave the conductors at the far end disconnected. Connect one conductor to the transmission line and excite the antenna on 40 meters. Then cut the antenna (both

conductors) at the far end until the lowest swr is obtained. After this, connect the other conductor at the far end to the 40-meter conductor element and cut about 16 feet from the far end.

Now switch back and forth between 40 and 20 meters, and trim the total antenna length (always leaving the two conductors shorted together at the far end). Trim the 20-meter stub to less than 16 feet from the far end. With care, lengths will be found that give a very low swr on both bands. If triband operation only is desired, the antenna can be operated as is. If 10-meter operation is also desired, cut the 20-meter stub side of the twinlead about 7 feet from the antenna base. Excite the antenna with ten-meter energy and try shorting positions between the upper portion and the 40-meter connector until minimum swr is obtained. Then cut the upper point of the 10-meter stub until a final swr minimum is attained.

This procedure tends to mess up the twinlead unless done with care. When trimming a conductor, the dielectric should not be cut away entirely. The conductor should be separated from the dielectric and then cut. Also, the proper shorting point between the two conductors can be found by pressing a pin through the dielectric to short the conductors. Only after the proper point is found should a jumper be soldered between the conductors. A proper ground connection is just as necessary for effective use of the twinlead antenna as it is for the fixed-station vertical.

fig. 2. Construction of a stub-switched antenna made from 300-ohm twinlead; complete tuneup procedure is outlined in the text.



summary

The stub decoupling method for multiband antennas is an effective, no-compromise method of automatic antenna bandswitching. Some designs require careful initial adjustments to establish initial dimensions, but the effort is well rewarded since no later maintenance work need be done. The basic idea of stub multiband operation can be used for any band where a suitable harmonic relationship exists.

The twinlead version of the antenna suggests the possibility of combining two such sections to form a multiband dipole antenna or inverted vee. Although this has not yet been tried, it would seem to be an extremely simple way to build an inexpensive multiband horizontal dipole for 40-10 meter coverage. No traps are required, and the stub arrangement should allow operation over a wider portion of each band with lower swr than is possible with high-Q traps.

ham radio



". . . 50 in a 35-mph zone. Besides that, you were QRMing my receiver."

glass semiconductors

Jim Ashe, W1EZT

Several newspaper, technical and scientific reports have appeared since last July about a new solid-state technology called "ovonics." Some of these reports suggest ovonics is a has-been before it ever arrived, and others say ovonics is a powerful new technology. I think we are going to hear a lot more about ovonics in the next few years, and possibly, eventually see ovonic IC's in communications gear.

glass semiconductors

The insulating properties of glass are well known. But not all "glasses" are glass, and glassy materials are apparently more organized in their structure than some recent reports suggest. The key to understanding what ovonics is all about is in the special almost-crystalline structure of glassy solids (see **fig. 1**).

At the left you'll see what I'll call a BS crystal; B *for big atoms and S for small* ones, in equal proportions. Now, looking at this we shortly observe that each B atom has four S atoms, and each S atom four B atoms, in its immediate vicinity. The entire assembly is arranged in neat ranks like the squares along and across a chess board.

At the right of **fig. 1** is a BS glass. It looks quite disorganized, but if you start counting you will see that practically every B atom has four S neighbors, and every S atom has four B atoms, just as in the crystalline state.

It turns out that the electronic behavior of semiconductors depends upon short-range order, within a few interatomic spacings of any given atom, and not upon long-range crystal-type order. If the BS crystal is an intrinsic semiconductor (will work without p or n doping) then the BS glass could be one also and offer fewer manufacturing problems. This is the way things have worked out in glassy-state physics.

For example, a glass laser rod three inches in diameter by four feet long does the same job as a crystal that would be nearly impossible to make. And once researchers understood some glassy state theory, engineers were able to develop an electronic conducting glass that carries an electron current in much the same way as

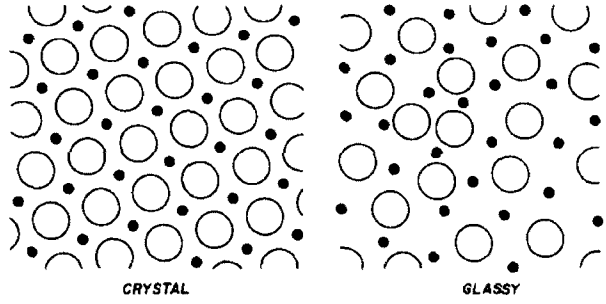


fig. 1. Glassy-state structure resembles crystal structure in that a given atom has about the same neighboring atoms in either case.

a piece of wire for use in image-orthicon tubes. Targets made of the new glass do not fail from electrolysis, so tube life is much better than earlier tubes using natural-glass structures.

ovons: the new devices

In 1962, Bell Telephone Labs constructed some simple glass semiconductor devices but they dropped this work in favor of the more familiar crystal physics semiconductors. This left the field open for Stanford Ovshinsky, who already had some strong patents and was working hard to improve the technology and build a company, Energy Conversion Devices.

Ovshinsky called his new semiconductors "ovons," and it seems likely this name will stick. While working hard, he has been very close-mouthed about his results until recently, and this has given his work something of a crankish appearance. The competition, perhaps slightly upset by thin-film ovons in transistor cases which appeared to be empty when opened, has failed to compete. Now Ovshinsky and Energy Conversion Devices seem to hold dominant patents in the new field.

An early ovon construction is shown in **fig. 2.** This is simply two graphite beads coated with some glass semiconductor material and held in mechanical contact. This package may be the origin of some early reports of erratic ovon performance. Later devices are assembled as shown in **fig. 3**—as thin films applied to an insulating substrate. It is reported that ovons can be made considerably smaller than transistors, and it appears this is because there are no multiple diffusion and electrical connection processes in manufacture.

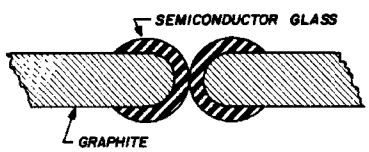


fig. 2. Mechanical construction of an early ovon. Improved thin-film ovon structure shown in fig. 3 to the right; the active material is applied without critical diffusion operations.

Published reports are not very clear on how ovons work. This may be Ovshinsky's close-mouthed policy again, making the competition work harder. Or it may be that details are not completely worked out yet. One suggested mechanism is a kind of solid-state lightning, in which an applied voltage leads to gradual warming of a channel between the two electrodes. This warming encourages current to flow, and a regenerative situation develops that leads to an abrupt breakdown. This is consistent with the input-output characteristic shown in **fig. 4,** but does not square with another fact.

That is, a minor adjustment in the glass chemistry results in an ovon that remembers which state it was in the last time power was applied to it. This is like a bistable that remembers if it was holding a zero or a one before it was turned off, and comes back on in the same state.

Ovons also show a much greater resistance than transistors to the degrading effects of nuclear radiation. As a result, the new technology is already being applied to satellite computers and other applications where environmental conditions are too harsh for transistor technology.

Circuits in ovonics tend to resemble those of two-terminal pnpn devices (see **fig. 5**). Here is a very simple relaxation oscillator that will generate an approximate sawtooth or a pulsed output. Other circuits use pairs of ovons in series as shown in **fig. 6.** Here, two 20-volt ovons in series are connected to a 35-volt supply. They do not break down until a five-volt pulse of either polarity is applied to the input terminal. This biases one ovon into conduction, and the other ovon follows.

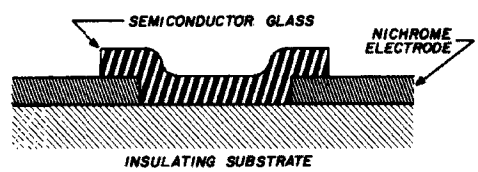


fig. 3.

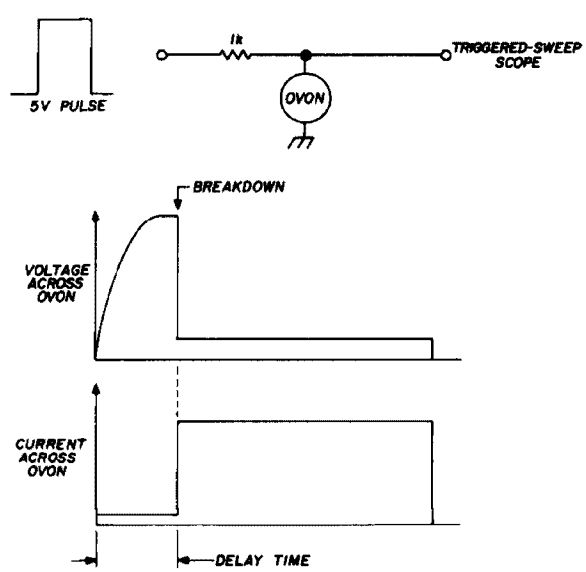


fig. 4. Ovon action when a square wave is applied.

the future of ovonics

Any estimate of a new technology's future must be at least 50% guess, but I think ovonics will eventually amount to something. I noted Stanford Ovshinsky seems to be playing his cards very close, seeking a long-term business advantage at a small price in short-term losses. I read that ovonic devices are actually in labo-

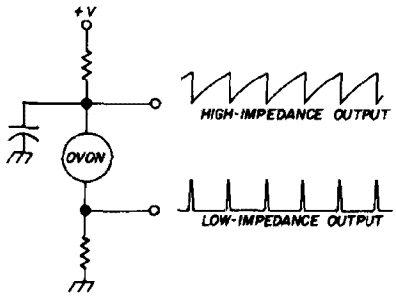


fig. 5. A simple ovon relaxation-oscillator.

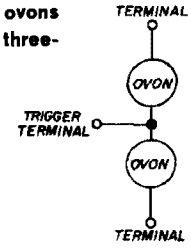
ratories, and people are thinking very hard about their applications.

For instance, the Johns Hopkins Applied Physics Laboratory has carried out extensive work in satellite computer design using ovons. And a reliable report that Energy Conversion Devices has been producing up to 150,000 ovons per day is also a convincing fact. Finally, there was a scientific paper describing the key principles thoroughly enough to enable a competent worker to make his own ovons. All this is too strong a basis for a trivial fad.

So what will this come to? Since ovons do not, at present, seem to have any linear-amplification capability, I think they will appear largely in computer applications over the next few years. But that is not as far from communications electronics as it used to be. An elaborate tv receiver, recently introduced in Europe's confused tv systems arena can receive color or black and white in any of the French, Russian, German or American systems of transmission. Priced at around $1,000, the receiver contains a considerable amount of computer information processing circuitry. Computer techniques have been used successfully in tv picture transmissions from Mars, and in the radar exploration of Venus. Finally, we have the class-D system of using switching circuits to develop audio power and manage power supply systems.

These considerations suggest there is a strong place for ovonics technology in coming transmitting and receiving gear—as computer-type circuits doing jobs now assigned to linear circuits. In transmitters, you may find tiny ovonic integrated circuits digitizing speech and image signals. And in receivers you may find inexpensive ovonic IC's doing computer analysis of incoming signals, in real time, paring off unwanted natural and manmade noise to achieve good reception at very poor signal-to-noise ratios. Finally, it appears very likely that ovonic technology will find applications in power handling circuits at lower costs and greater reliability than present crystal semiconductor technology can offer.

fig. 6. Using a pair of ovons in series to obtain a three-terminal active device.



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E. Evans, "A Feasibility Study of the Application of Amorphous Semiconductors to Radiation Hardening of Electronic Systems," $3.00 from Clearinghouse, Sill Building, 5285 Port Royal Road, Springfield, Virginia 22151. Report No. PATR 3698, AD No. 670-949.

ham radio

a 40-meter bobtail curtain array

A modified three-element vertical broadside antenna that will more than double your radiated power

George Cousins, VE1TG, R. R. 2, Box 18, Lower Sackville, Nova Scotia

Want to have a lot of puzzled hams throwing questions at you over the air? It's easy. Just say you're using a bobtail-curtain array for an antenna. This is guaranteed to keep you busy for an entire evening, especially if you're laying down a healthy signal. This is what happens to me when I use this antenna on 40 meters. Its name gets everyone so worked up that I've stopped being honest and just say I'm using a vertical.

Despite all the confusion, this is an antenna worth knowing more about. It has the advantages of simple construction and tuning, low cost, and gain at low radiation angles. It's an excellent antenna for any of the low-frequency bands, and performs especially well on 40 and 80 meters. The only major disadvantages are height and area requirements and the necessity for an antenna tuner.

description

Over-all dimensions are shown in **fig. 1**. It is an offshoot of the three-element vertical broadside array. In the classical version, three elements are fed in phase with equal currents, and the elements are spaced one-half wavelength apart. This arrangement theoretically has no radiation off the ends because of the phase relationship existing between elements. In the bobtail array, there is some high angle radiation off the ends. This occurs because of imperfect cancellation due to the flat top portion. So there's a small compromise. Compared with the three-element co-phased vertical array, the bobtail's front-to-side discrimination is somewhat degraded. But its broadside gain is still pretty good—7 to 10 dB over a reference dipole.

The height requirement for 80 meters will be about 70 feet, but for 40 meters it's only about 35 or 40 feet. The flat top portion of the 40-meter version is just an 80-meter dipole with no center insulator, so if you have room for this you can build the bobtail.

construction

There are several ways to go about construction. If you can get the two ends supported about 40 feet or more above ground, there will be just about enough sag in the flat top to allow the center wire to hang straight down to the tuner. However, I didn't have quite enough height at the ends, so I used a different approach. My center element is actually a 40-meter ground plane, made of three 12-foot lengths of aluminum tubing, telescoped together and adjusted to approximately 34 feet over-all. The joints are split and clamped tightly with stainless-steel hose clamps well coated with zinc chromate to prevent oxidation.

The element is mounted on an insulator originally used for whip antennas on tanks (bought through surplus channels). This insulator is mounted on top of an adjustable tripod mount (also surplus), and the whole element is guyed with three sets of three guys each, made of nylon cord. Bear in mind, when choosing the insulator, that rf voltage is high at this point.

The two half-wave phasing lines are soldered to heavy lugs and bolted to the top of the center element. At each end of the phasing wires, another vertical wire is connected, made of number 10 or number 12 wire. The ends of the flat top are then pulled up to full height so the verticals hang straight down. A little compromise will do no harm. I've used the antenna with the two end elements almost 45 degrees with respect to ground, and results have been just as good.

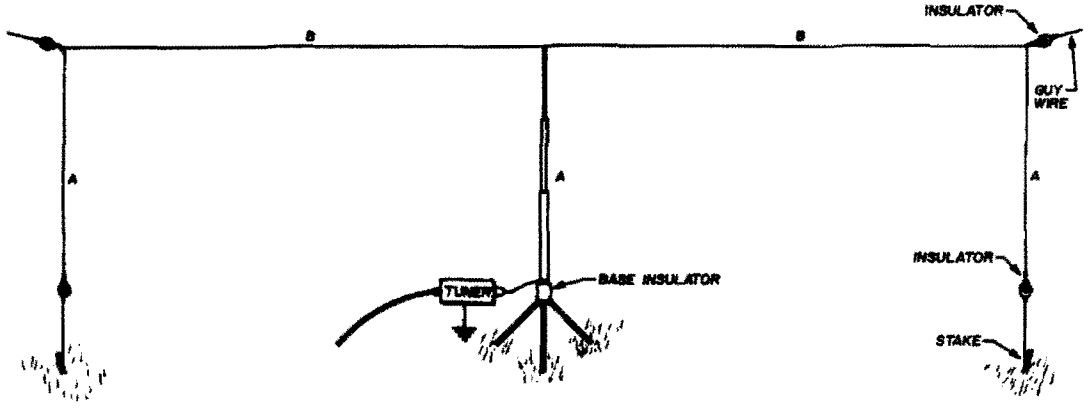


fig. 1. A bobtail-curtain antenna for the lower amateur bands; dimensions in the table are in feet.

Band	A	B
80	66	132
40	33	66
20	16.5	33

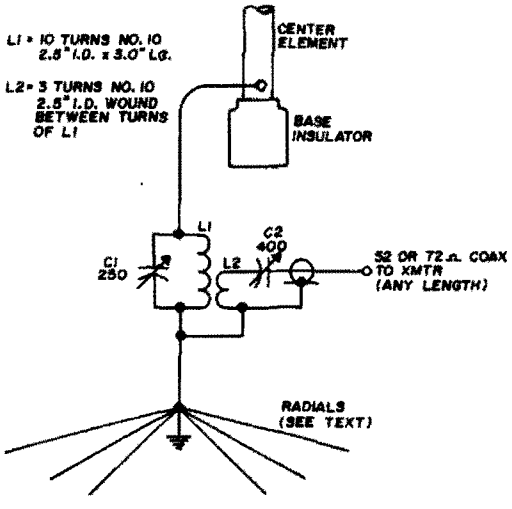


fig. 2. Tuning unit for the 40-meter array. For 20-meter operation, L1 and C1 should be approximately one-half the values shown here; C2 is the same.

antenna tuner and adjustments

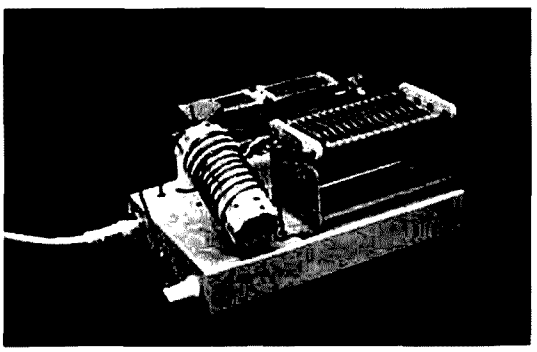
The antenna has a high input impedance, so an antenna tuner must be used. **Fig. 2** shows the circuit, and the photos illustrate tuners used for 40- and 20-meter versions. The coil for the 20-meter tuner is a B&W BEL-150. For 40 meters, the coil and link are wound on the ceram-

ic form from a BC-375 tuning unit. The small capacitor tunes out reactance in the transmission line, which in my case is over 100 feet long.

Tuneup consists of adjusting both capacitors to obtain minimum standing-wave ratio. I made use of my ever-patient wife and my children's walkie-talkies to overcome the distance problem from tuner to shack. Only a few minutes were required to bring the standing-wave ratio to almost 1:1, and I have had equally easy tuneup with two other bobtails, which I built in the past for 40 and 20 meters.

To avoid hanging the antenna so the driven element would have to come into the shack to the tuner, I left the tuner at the antenna base and made it waterproof. After tuneup was completed, I enclosed the tuner in a box made from clear sheets of plexiglass. This makes an excel-

Tuner for the 40-meter bobtail.

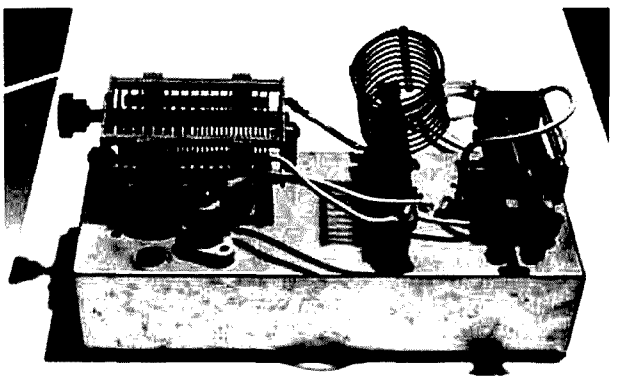


lent weatherproof container, and the components can be seen at a glance.

ground system

Undoubtedly if you researched enough antenna handbooks, you'd find a lot of information on the bobtail, but I have found very little. It appears that an elaborate ground is not necessary, and a 4- to 6-foot ground rod seems ample,[1] connected to the tuner through a flexible piece of braid or heavy wire.

I still had the eight radials used for the center element when it was just a ground plane, so I decided to bury the radials directly under the center element. If I were to say this made a fantastic difference or even any difference for that matter, it wouldn't be true. However, it seems logical that a good radial ground system should be just as effective for this antenna as for any other vertical array, so I like to think that those fine reports on 40-meter DX contacts are just a bit better because of the radials. I've been tempted to install radials under the outer elements also, but so far I haven't done so.



Tuner for the 20-meter bobtail is similar to 40-meter unit but with smaller components.

results

The antenna tunes broadly, and the standing-wave ratio remains reasonable over the whole band. My broadside pattern is beamed northeast—southwest, and the antenna has produced many good comments from European, Near-East and Pacific stations. Results in North America aren't spectacular, because the radiation angle is low. It's a DX antenna, and doesn't really start to perform until the distance is greater than 2000 miles or so. After that, it's a great antenna! It would be interesting to hear from others who may be using this antenna so I could compare notes.

references

1. *The Radio Handbook*, Editors and Engineers, Santa Barbara, California, 1959 edition, pp. 472-473.

ham radio

propagation

predictions for july

Victor R. Frank, WB6KAP, 12450 Skyline Boulevard, Woodside, California 94062

March and April were months of frustration for many six-meter operators. I mentioned in my September 1968 column that **then** was the time to send equipment and line up 50-MHz activity in the South Pacific; I pointed out that places like Easter Island, Pitcairn, Tahiti, Samoa and Fiji could be worked **if** amateurs there were active on 50 MHz. My predictions were borne out, as amateurs in the Caribbean, Mexico and Gulf States will testify.

Although few contacts have resulted, ZK1AA's beacon has been heard in these places and others during February, March and April. Perhaps the most widespread, strongest opening was between 2100 and 0038 gmt on April 11 to Louisiana, Mississippi and Alabama. K5AGI again heard the beacon weakly between 1913 and 2010 gmt on the 12th. HI8XDS, VP2AD, XE1GE and XE1PY also report hearing the beacon frequently. The first I heard the beacon this year was on April 17 between 0500 and 0645 gmt by nocturnal TE. ZK1AA was copied in Northern California and Washington (by W7FN) again on the 19th (0429-0550 gmt), the 20th (0712-0734 gmt) and the 21st (0520-0812 gmt). The beacon was also heard in Northern California on April 24, 27 and 28. It was heard in Southern California on the 25th and 28th.

ZK1AA indicates that his receiving system was improved considerably on April 19th. Also, W6ABN sent him a four-element beam. I worked him on April 21st at 0535 gmt for his first W QSO. He reported working ZF1AA on the 19th at 2030 gmt and XE1PY at 2040 gmt as well as about twenty JA's. During the last week of April he worked W6ABN, W6YDF, K6QEH and KH6GMV.

To work him, weekends and weeknights, attract his attention with a string of dots swept across his frequency and call him about 500 Hz higher. His transmitter is remote controlled from his home. As of this writing his receiving installation is somewhat under par and amateurs are advised to use cw rather than ssb when calling.

Also, 5W1AR in Western Samoa is scheduled to have his beacon on; it's of the simplex variety so he is not able to listen while transmitting.

Openings between the West Coast and South America were reported on March 16 and 30, and April 5, 8, 9, 10 and 11. Backscatter and sidescatter openings from the West Coast were reported on March 30 (W3, W5 and W6), April 1 (W4, VP2MJ), April 4 (HI8XDS), April 5 and 6 (W6). Other F2-layer openings include March 24, when KH6NS, KH6EQF and KH6GHC worked from VE7, W6, W7 and W5; March 25, when KH6EEM and KH6EQF worked 5W1AR and KH6EQF worked K7HER/KG6; and April 11 when XE1PY worked ZF1AA.

One of the rare openings of the year was the aurora of March 23 to 24 which extended as far South as the San Francisco Bay area (I worked K7ZIR in Portland on 144 MHz).

W7FN, Seattle, reports aurora as early as 1400 pst; he worked KL7GLL, all 7th District states except Nevada and Arizona and as far east as North Dakota on 50 MHz.

I believe that these 50-MHz openings were due to the general increase in peak solar activity that began about February 20, combined with seasonal effects. There are some mysteries, however, like why the ZE1AZG beacon (50.05 MHz) was not heard in the Americas—the path is similar to the ZK1AA—Caribbean path. Also, why was nocturnal TE so late getting to the Bay area?

sunspot numbers

I read WAØIQN's predictions in **QST** about increased solar activity in upcoming months with great interest. Don even suggested that the sunspot numbers might go as high as 200 this year. Upon investigation, however, I do not see such an optimistic view—here's why. There are more than one "brand" of sunspot numbers in use, a fact I neglected to mention last month when I gave daily sunspot number counts between 200 and 300 for February 20 to 27, 1969.

These numbers came from preliminary ESSA reports of solar-geophysical activity. They are perfectly valid sunspot numbers, but they were taken at the ESSA-Boulder or Sacramento Peak (Colorado) observatories—not at Zurich. You may remember that in my November 1968 column on sunspot numbers, I mentioned that a correction factor had to be applied to observations taken at observatories other than Zurich to account for differences in observation.

Actually it's not quite that simple, and most ionospheric forecasters stick with the sunspot numbers measured at the Swiss Federal Observatory at Zurich. The difference between the ESSA and Zurich sunspot numbers can be quite substantial.

I have plotted in **fig. 1** the provisional daily sunspot numbers observed at Zurich vs those observed by ESSA between January 3 and February 28, 1969. You may ask, "Is that really the same sun they're observing?" Yes, it is, with an 8-hour time difference, but it appears that ESSA can frequently count 50 percent more spots than Zurich.

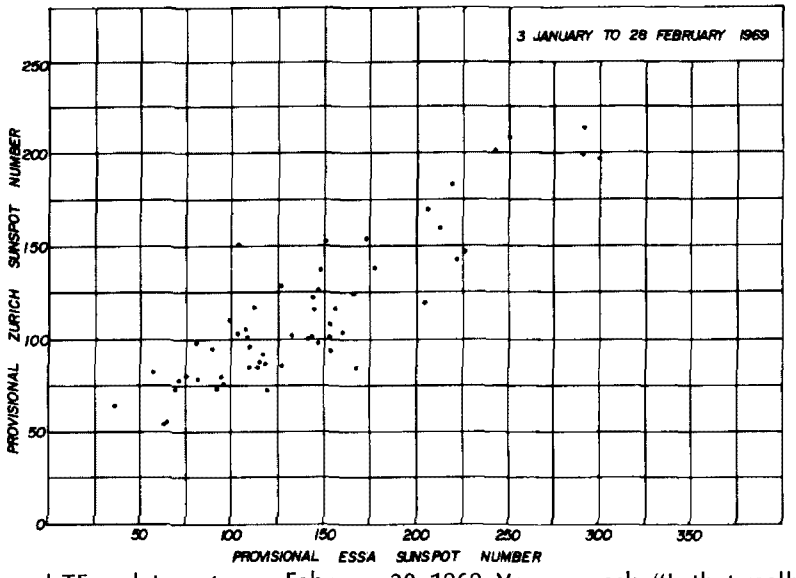


fig. 1. A scatter plot of daily sunspot numbers measured by the Swiss Federal Observatory vs those measured by ESSA for the period January 3 to February 28, 1969.

With this in mind, I'll admit that if solar activity continues its February/March rise, monthly average ESSA sunspot numbers of 200 are not inconceivable, but I doubt that the smoothed Zurich sunspot number will exceed 120 (this spring or summer). I am basing my predictions on a smoothed Zurich sunspot number of 100. The highest smoothed sunspot number for cycle 20 to August 1968 is 107.6, which occurred in May 1968.

In view of the present uncertainty in the course of cycle 20, last year's ionospheric observations are as good an indicator of probable ionospheric conditions this year as any. Thus, I scaled some July 1968 ionograms from Pt. Arguello, California to determine the absolute muf (somewhat greater than the 4000 km muf) for the hours of 0500, 1300 and 2100 local time (pst).

fig. 2. Daily F2-layer absolute muf's scaled from Pt. Arguello ionograms, daily sunspot numbers and magnetic activity indexes during July 1968.

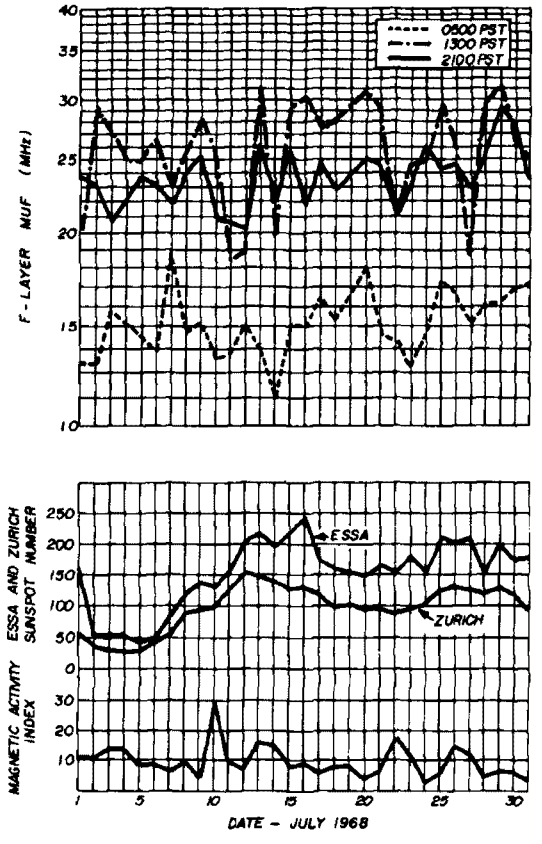
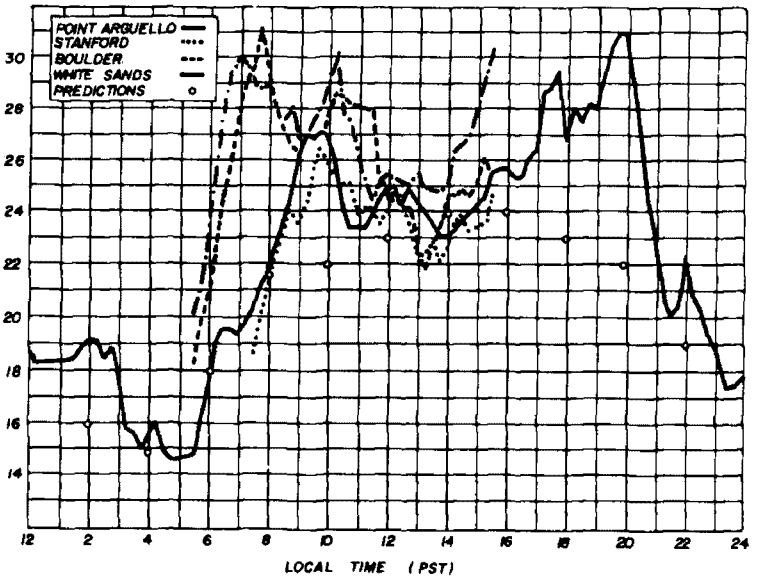


fig. 3. Absolute muf's scaled from July 10, 1968 ionograms from four stations in the Western United States.



The day-to-day variation throughout the month, shown in **fig. 2,** is quite substantial. Based on these measurements, ten meters should open to control point latitudes of 35.5° N. at 1300 local time more than 33 percent of the days of the month. Fifteen meters should be open more than 80 percent of the days at 1300 and 2100 local time. Twenty meters should be open more than 75 percent of the days of the month through the predawn minimum at 0500 local time.

Plotted along with the Pt. Arguello muf's are the daily sunspot numbers of ESSA and Zurich and the magnetic activity index. The most important solar event of July 1968 was an importance-3B flare on the 8th which produced intense vhf/uhf radio noise bursts and moderate x-ray emission and SID (sudden ionospheric disturbance). It was followed by particle emission reaching the earth 29 hours thereafter (1400 pst on the 9th); this caused a minor magnetic disturbance which continued for some 30 hours and depressed ionospheric critical frequencies as late as the evening of the 11th.

Another magnetic disturbance was produced as this spot group passed the solar central meridian commencing on the 13th (8 am pst). Note the rise in muf's on the first day of the disturbance and the depression on the following day and night. Muf's were also depressed on the 22nd and 17th due to minor magnetic disturbances.

Since the normal summer daytime muf and evening muf's are not very much higher than the ten and fifteen meter bands, any reduction of muf's due to minor magnetic storms may be more noticeable than during winter months. Frequently during these storms sporadic-E will be present, allowing contacts to 1400 miles or so, but no DX. Also, in addition to decreased F2-layer critical frequencies, the F2-layer virtual heights

may rise to over 500 km indicating possible long-skip, single-hop paths as long as 3000 to 3600 miles.

When considering paths of this length, however, the variation of ionospheric parameters along the path becomes important. Ionospheric soundings taken at the oblique path midpoint may no longer furnish a sufficient model of the ionosphere to explain all the observed phenomena.

To demonstrate the variability of the ionosphere in distances as short as 300 to 1000 miles, I have scaled daytime ionograms taken on July 19, 1968 at four stations: Stanford, California, Pt. Arguello, California, Boulder, Colorado and White Sands, New Mexico. As I pointed out in the August 1968 column, prediction methods assume that, over a limited range of longitudes, the muf contours are fixed in space relative to the sun and the earth rotates underneath. You would expect the muf-vs-time curves of the California stations to be offset one hour from the other two. You would also expect, from the pronounced north-south gradient of ionization predicted that White Sands would have the highest muf's, followed by Pt. Arguello, Stanford and Boulder in that order.

Fig. 3 shows the scaled absolute muf vs time (pst) for each station along with ESSA predicted muf's for Pt. Arguello. Changes taking place simultaneously at all stations would likely be due to changes in the incident solar ultraviolet flux. Changes that occur with time delays from north to south between the stations may be due to large traveling ionospheric disturbances or transport of ionization. Changes that occur with time delays of one hour east to west are likely due to the normal diurnal variation. The picture is not clear from just one days' data—but it shows what differences you might expect from the standard predictions.

Between 1030 and 1130, the normal north-south ionization gradient was reversed at both pairs of stations. The eastern pair of stations held the edge on muf's (shifting to local times) until late afternoon. According to predictions (**fig. 4**) the daytime F2-layer 4000-km muf at Boulder should have been 3 MHz lower than that at White Sands.

One of the major problems facing ionospheric scientists in interpreting long distance propagation is the development of suitably accurate models of the ionosphere. The four stations at Stanford, Pt. Arguello, Boulder and White Sands are quite closely spaced compared to other sounders. Could you imagine ESSA trying to forecast weather with a network of observing stations spaced this far apart? There are still frontiers to be conquered with the vertical-incidence sounder, but after 33 years it is still the prime tool of the ionospheric physicist.

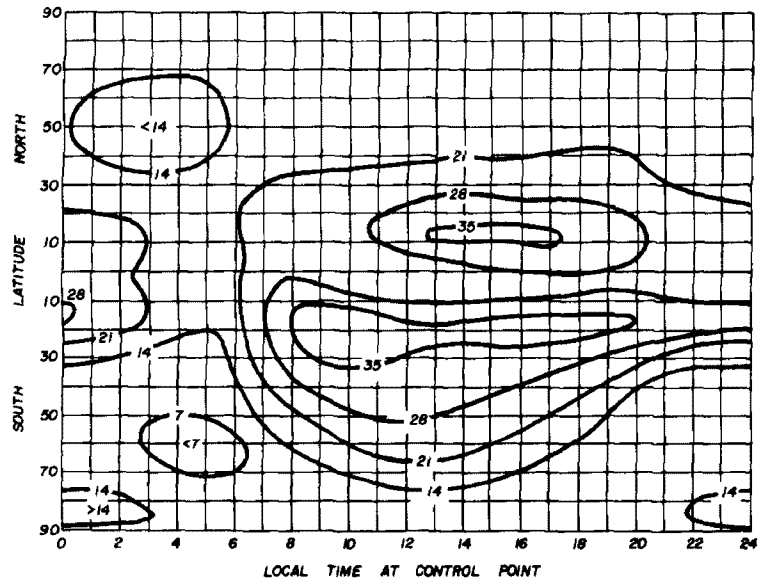


fig. 4. Predicted median muf (4000) F2 for July 1969 centered on longitude 105° W.

propagation during the month

Summer is still with us, and propagation during the first half of July will be very similar to that during June. The same maximum range charts may be used. Some idea of occurrence of sporadic-E may be gained from the propagation column in the June issue of **ham radio.**

Table 1 lists some specific predictions for Northern California. They may be of use to other areas the same distance and bearing from mid-latitude United States. San Francisco to Anchorage is equivalent to Washington,

Table 1. Times (pst) of predicted band openings to San Francisco for July 1969.

From	Distance	Bearing	40 Meters	20 Meters	15 Meters	10 Meters
KL7, Anchorage, Alaska*	1960	332	2000-0700	0630-0200	nil	nil
3W8, Saigon, Vietnam	7650	306	0200-0700	2120-0400 0600-1030 (0100-0430) (1500-1900)	1930-2030 0930-1100 (1500-2200) (0815-1130)	(1700-1730)
KR6, Okinawa, Ryukyus	5960	303	0030-0700	1930-0400 0600-1130	1900-2030 (1430-2200) (0600-1130)	(1530-1730)
JA3, Tokyo, Japan	5020	303	0000-0700	1830-0400 0600-1130	1900-2030 (1330-2200) (0615-1130)	nil
KG6, Guam, Marianas*	5620	282	2330-0720	1930-1100	1140-2200	nil
KX5, Kwajelein*	4650	265	2200-0730	1730-1120	1030-0200	1930-2030
VK6, Perth, W. Australia	9100	259	0320-0600	2100-1030 (1500-2300)	1400-0200 (1730-1830)	1800-2100
VK3, Melbourne, E. Australia*	7000	245	2300-0630	1900-0900	1130-0130	1400-2100
KC4, Antarctica*	7590	180	1900-0300	0800-0900	0930-1600	1130-1400
LU, Buenos Aires*	6700	128	1820-0200	0200-0640 1420-0030	0500-2200	1100-1830
KZ5, Canal Zone	3280	116	1740-0320	24 hours	0600-2130	1200-1700
ZS1, Capetown, S. Africa	10150	76	1900-2100 #	2130-000 (2300-0800)	0900-1030 (2015-0200)	nil
DL, Berlin, W. Germany	5540	25	1900-2100 #	0530-0030 (0230-0700) (1530-1830)	(0700-1600) (2100-0200)	(1300-1230)
W3, East Coast USA*	2420	74	1720-0300	0700-2330	1500-1900 **	nil**

( ) Time of long path opening (opposite bearing)

* No long path calculations made for these paths

\# Difficult circuit—high path losses

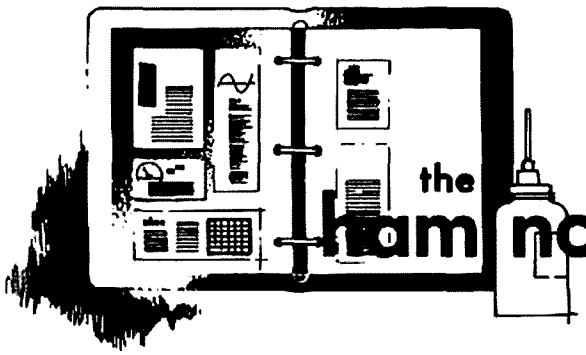
nil Muf not supposed to rise this high—possible opening during middle of lower band opening on less than half the days of the month

** Fifteen may open as early as 0600 on days with high muf's, ten may open a half-hour later and close a half-hour earlier than fifteen. Double-hop sporadic-E may occur anytime.

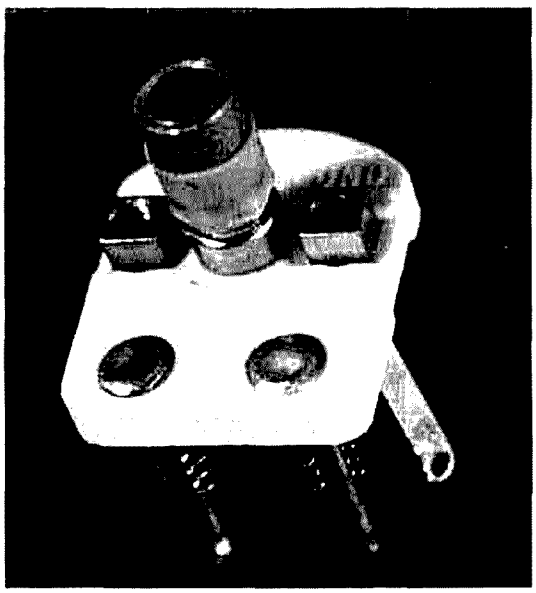
D. C. to Yellowknife N.W.T., SF to Saigon is equivalent to Kansas City to Guam and SF to Melbourne is equivalent to St. Louis to Auckland, New Zealand. Times are your local solar time.

Late note: as of May 20, ZK1AA had over 50 contacts in less than two months—all in the Northern Hemisphere. His beacon is operating daily from 1730-0930 gmt. 5W1AR, Western Samoa, now has a beacon on 50.105 MHz operating 1900-2400 gmt and 0400-0600 gmt; he is bothered by interference from television on American Samoa that comes on the air at 0600 gmt.

ham radio



adding shaft to apc trimmer



Ever faced with that very frustrating situation where the only variable capacitor in the junk box with the required value is an APC trimmer—and you need one with a shaft? Although several solutions have been presented in the past for this problem, the following one is easy, effective, and thus, my favorite.

Slip a piece of shrinkable tubing over the hexagonal rotor extension, and trim to extend at least 1/8-inch. Insert a suitable length of 1/4-inch diameter metal tubing into the open end; apply sufficient heat to shrink the tubing. Now, gently slide the shrinkable tubing and shaft extension back off the hexagonal rotor extension. Note that it comes off easily but retains the hex shape. Put epoxy on the exposed inner surface of the shrinkable tubing and replace on the rotor extension. Now pour epoxy through the center of the metal tubing, making sure it flows into the original screwdriver slot and that no air pockets are left.

Let it stand for at least twenty-four hours before attempting to turn the shaft. You now have an APC trimmer with a rugged shaft extension; add a knob and you're in business.

V. M. Scott, Jr., W1ETT

the multi-box

Anytime you put a piece of audio gear on the bench there is the problem of mixing and matching the variety of connectors used in current equipment. My little multi-box has proven invaluable when working on such gear. In its present form it consists of three types of jacks mounted in pairs on a 2 1/4 x 2 1/4 x 5 minibox. A terminal strip mounted on the left-hand end of the box accepts pigtail connections. The jacks and terminal strip are all wired in parallel so that any signal fed into one jack is available at any other jack. The three types of jacks on my personal version of the multi-box are standard phone jacks, mini jacks (á la transistor-radio practice) and RCA phono jacks. The RCA phono jacks were mounted with the aid of pop rivets; this is a very neat way to mount this style jack in any gear you may construct.

Al Joffe, W3KBM

simple scope calibrator

An accurate scope calibrator is a very useful tool where an oscilloscope is frequently used. It eliminates calculations due to scope gain controls or a 10:1 probe since the tip of the probe can be inserted into the calibrator output jack for comparisons with the signal voltage.

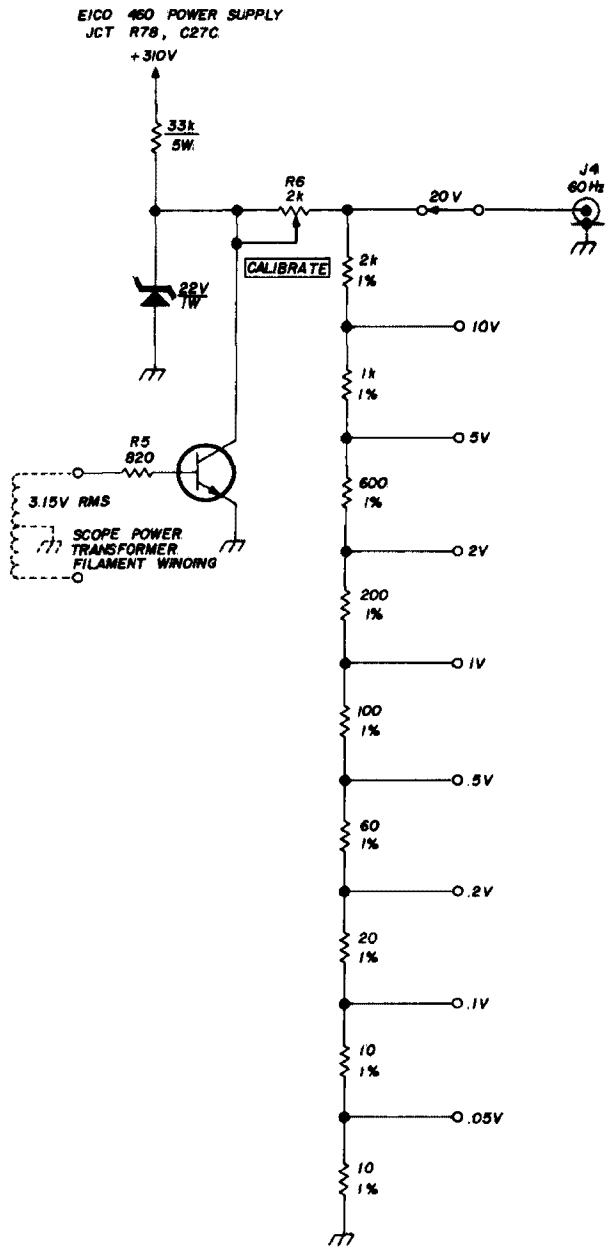
The calibrator shown in **fig. 1** was built in one afternoon and added to an Eico 460 oscilloscope. The same circuit could be added to other inexpensive scopes or built as an output accessory in a minibox. It's simple, the cost is low and the added current drain on the scope power supply is negligible.

The calibrator voltage is derived from a 22-volt zener diode. It is chopped at 60 Hz by the transistor and the resulting 60-Hz square wave is trimmed to exactly 20 volts and applied across a precision resistor voltage divider. The calibrator can be adjusted with an accurate dc volt-meter by disconnecting the transistor collector and adjusting for 20 volts at the top of the divider.

To install the calibrator in the Eico 460 scope, remove the pilot light, X1, and resistors R5 and R44. The pilot light is a filament tie point, so the new wiring must take this into account. The best procedure is to run new wires from TB2 along the path of the old leads. Wiring to the phasing control (R45) is checked for continuity, and the scale dimmer (R80) is wired directly to TB2. The pilot light mounting hole is used for the calibrator rotary switch, and the "60-Hz" binding post is used for calibrator output.

The rotary calibrator switch is completely assembled before it's put into the scope; 1% resistors are used in the divider for maximum accuracy. After mounting the rotary switch, the zener and chopper transistor are mounted on an additional two-terminal tie strip mounted over TB4 and near XV6. The calibrate potentiometer (R6) is rotated 180 degrees in its hole and wired into the circuit as a voltage trimmer. The 820-ohm resistor, R5, is re-used in the base circuit of the transistor.

The transistor may generate voltage spikes that will crosstalk into the scope input unless the new wiring is dressed close to the panel where it's connected to the 60-Hz post. Also, the divider ground lead should not be connected to the scope input ground post—return it to the point where the transistor emitter is grounded near XV6. Nearly any npn transistor will work as the chopper; I tried surplus 2N706's and 2N1302's among others.



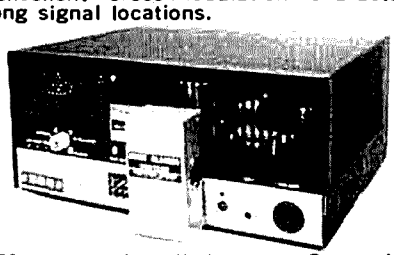
Bert Kelley, K4EEU

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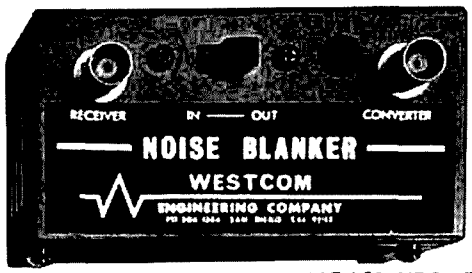
TNB-250 shown installed on a Swan 250 at accessory socket location.

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Refer to the New Products column of the August '68 issue of Ham Radio Magazine for additional information on the TNB Noise Blanker or write for technical brochure.

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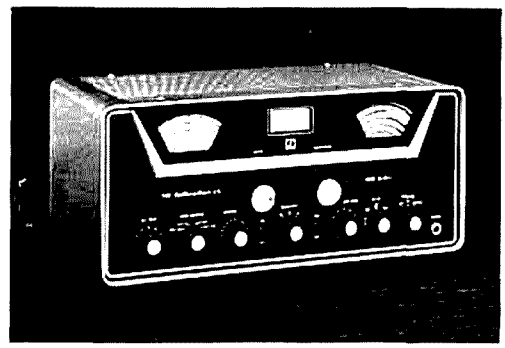
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new products

communications receiver



Hallicrafters has just announced an advanced version of their popular SX-122—a dual-conversion receiver providing general coverage am, cw and ssb reception. The new receiver, the SX-122A, covers standard broadcast (540 to 1600 kHz) plus 1750 kHz to 34 MHz in three bands. Bandspread is calibrated for 80, 40, 20, 15 and 10 meters. The SX-122A features extensive temperature compensation, voltage regulation and a crystal-controlled second-conversion oscillator for improved frequency stability. The dual dials provide fast and easy dial setting plus smooth flywheel tuning.

Other features of the SX-122A are a precisely tuned and tracked rf stage, variable selectivity (500 - 2500 Hz and 5 kHz), antenna trimmer, separte rf gain control, automatic noise limiter, s-meter, provision for an optional front-panel-controlled 100-kHz crystal calibrator and balanced or unbalanced antenna inputs. $350 from your local distributor, or write to the Hallicrafters Company, 600 Hicks Road, Rolling Meadow, Illinois 60008.

gate-protected dual-gate mosfet

RCA has just announced a new gate-protected dual-gate mosfet—the 40673—that has essentially the same electrical characteristics as 3N140 and is interchangeable with it. Gate protection is provided by back-to-back zener diodes which limit the voltage excursion of the gates to ±10 volts; by adding this protection from static burnout, the gate capacitances are increased about 1.5 pF. Noise figure of the basic device increases 0.2 dB, and power gain decreases 0.2 dB at 200 MHz; both changes are considered negligible. The price of the new RCA 40673 is about $2.00 in small quantities; since this is a brand new transistor, it may take awhile to reach the shelves of your local distributor, although he should be able to special order them for you. According to RCA, more gate-protected mosfet's will be available in the next few months.

dipole antenna kit

The new Mosley Electronics DIV-80 dipole antenna kit is designed to be constructed horizontally or as an inverted-V on any single amateur band from 10 to 80 meters. The amateur is supplied with all the necessary parts plus easy-to-follow instructions to determine the length required for specific frequencies within a given band. The DIV-80 is rated at 1000 watts a-m or CW and 2000 watts PEP ssb.

When assembled as a horizontal dipole, the antenna is raised between 35 and 65 feet above the ground and is approximately 135 feet long on 80 meters. When used as an inverted-V, the center support point is 35 to 65 feet above ground with the ends of the antenna insulated about 10 feet above the ground. The DIV-80 kit includes 140 feet of copperweld wire, a high-strength Mosley dipole connector, high-grade ceramic end insulators, all necessary hardware and complete instructions. For more information, write to Mosley Electronics, Inc., 4610 N. Lindbergh Boulevard, Bridgeton, Missouri 63042.

RCA has all-new FCC commercial license training

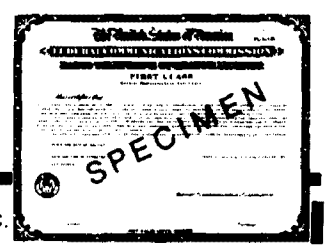
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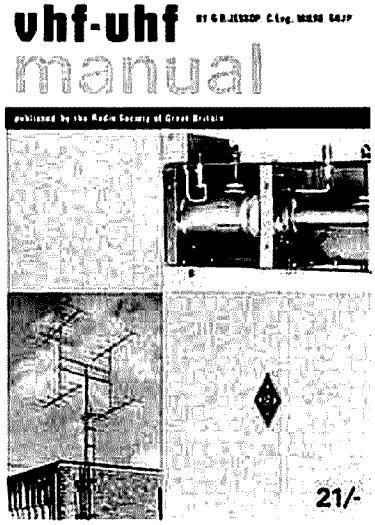
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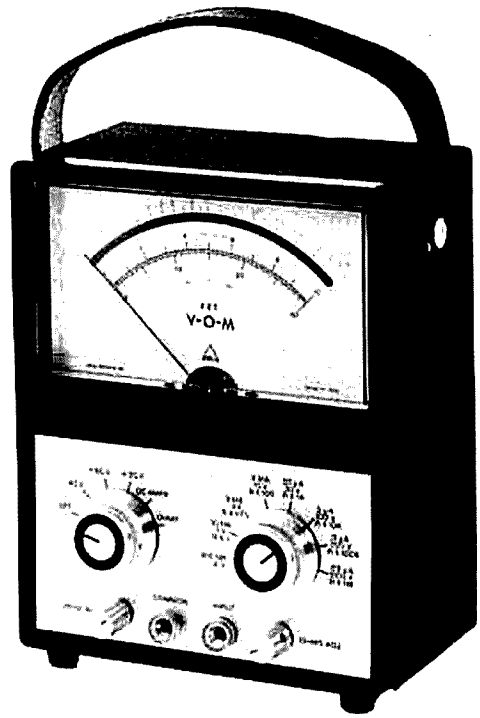
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fet vom



The newest of a series of precision electronic instruments has been announced by Delta Products. Their new model 3000 fet volt-ohmmeter features ac and dc ranges from 300 millivolts to 1000 volts full scale; the instrument reads down to 10 mV. Current ranges are from 0.03 μA to 300 mA, and resistance ranges run from 10 ohms to 20 megohms center scale. Dc accuracy is ±2% of full scale and ac accuracy is ±3% of full scale below 500 Hz.

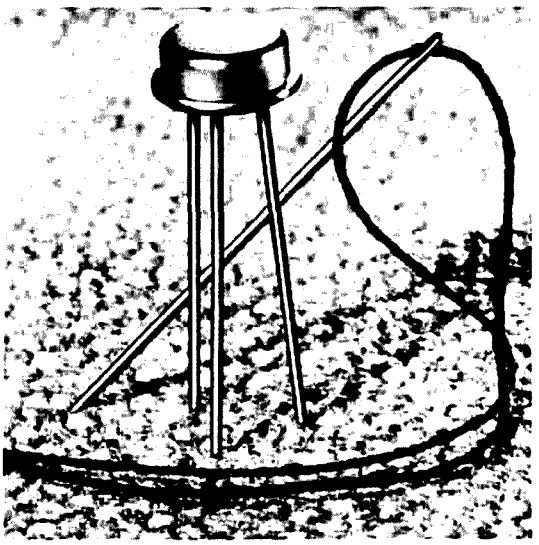
The model 3000 has an fet input stage with current regulator for 10-megohm input resistance; an integrated-circuit operational amplifier is used for extreme accuracy. The meter is fully compensated for low zero drift and has built-in voltage clippers for input stage protection. Special features include a mirror-scale meter movement, enclosed switches, and ten-turn potentiometers for zeroing and ohms adjust.

Available in kit form with preselected components for the feedback network to eliminate all calibration, $59.95 ($74.95 completely assembled and tested) from Delta Products, Inc., Department 147, Post Office Box 1147, Grand Junction, Colorado 81501.

solid-state tv camera modules

If you're interested in building your own vidicon-type tv camera or updating existing cameras, you'll be interested in the new line of solid-state modules announced by ATV Research. A complete camera can be built in one evening by using these wired and tested modules; and according to the manufacturer, no previous tv knowledge or special test equipment is required. The modules presently available include a video module, vertical-sweep module, horizontal-sweep module, rf modulated oscillator module and high-voltage dc-dc converter. Prices range from $10 to $20, and units are available for immediate shipment. For more information, write to ATV Research, 13th and Broadway, Dakota City, Nebraska 68731.

millisecond fuse



Beckman Instruments has just introduced a new high-speed fuse designed specifically to protect integrated circuits and power transistors. The series 817 fuse is inexpensive and available with ratings of 0.5, 0.75 and 1.5 amperes. All three units will blow as fast as 50 microseconds. The miniature fuses are housed in a TO-46 can. For more information, write to Technical Information Section, Helipot Division, Beckman Instruments, Inc., 2500 Harbor Boulevard, Fullerton, California 92634.

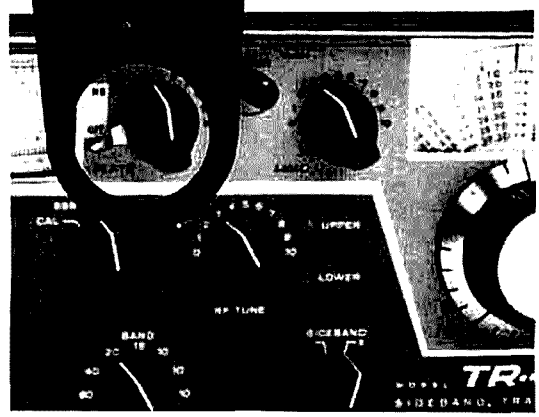
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Installation of the kit is about a two hour job for the competent technician only, requiring the usual hand tools, plus soldering iron and electric drill. Factory installation, $15 plus shipping.

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short circuits

silver plating

In the article on silver plating in the December 1968 issue of ham radio, it was stated that immersion plating is also known as electroless plating. This is an error. Immersion "plating" only puts on a surface film, then ceases to deposit. Electroless plating continues to deposit for the life of the solution.

220 mosfet converter

The 470-ohm isolation resistor in fig. 1, page 30 of the January 1969 issue does not isolate the hot end of L2—the hot end of L2 should be connected to the junction of the 470-ohm resistor and the 500-pF buttom mica bypass capacitor.

solid-state BC348

Several errors appeared in the circuit diagrams of this article on the solid-state BC348 which appeared in the February 1969 issue. The squelch switch, as drawn, shorts out the agc—the switch should be turned around so the arm is connected to the base of the 2N2924. In the negative 6.3-volt regulator, the collector of the 2N3638 should run directly to the base of the 2N1132/2N3638. In the positive 12.5-volt regulator, the 2N3766 driver should be a 2N3642 or 2N697; the main regulator is a 2N3766.

In the text, it was stated that the gain control range was 40 dB. Actually, for three stages, range is more like 140 dB! Fairchild Semiconductor no longer markets the 2N4122—use 2N4121's instead. Another solution, offered by W1OOP, is to use Motorola 2N3906's for all pnp types except the negative 6.3-volt regulator. For all npn types except the above mentioned 2N3642/2N697, use 2N3904's. Price is 82c each.

repeater identifier

In the power supply circuit in fig. 1, page 19 of the April 1969 issue, the plus 4.5-volt supply should be connected to the emitter of the transistor, not the base—emitter is lifted off ground of course. Also, in fig 5, page 22, the 1.8 k resistor from pin 5 of IC1 should be connected to the collector of Q23, not the emitter.

ham radio

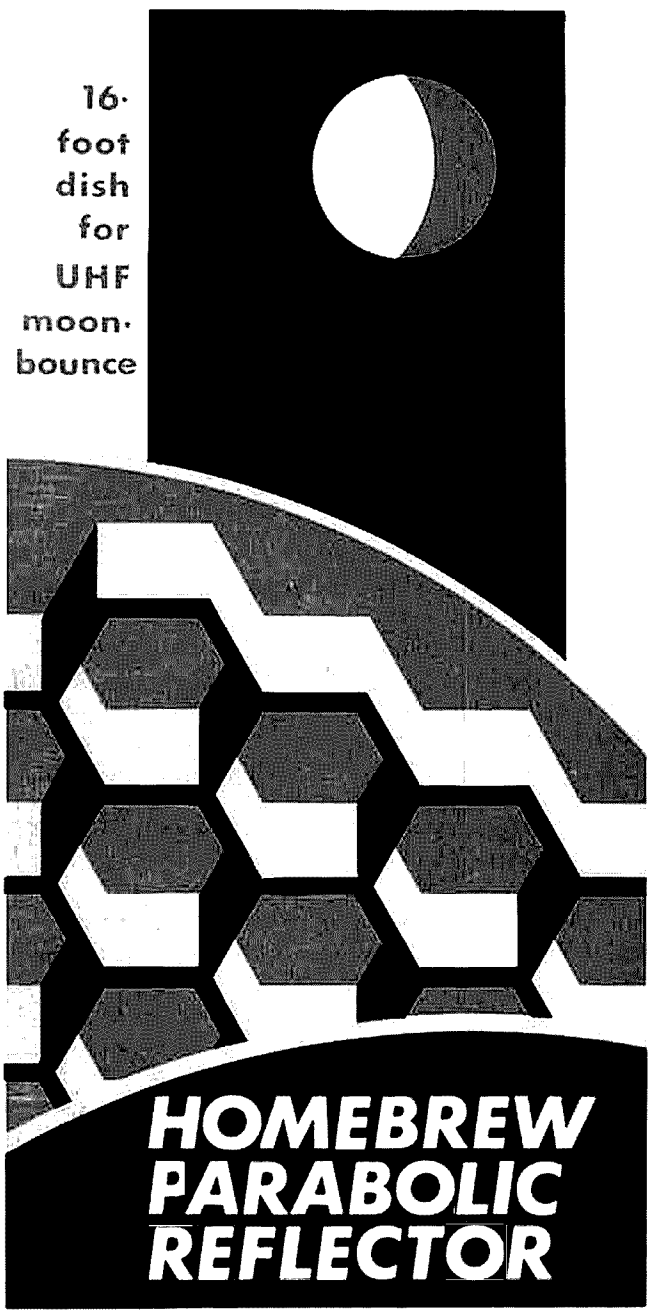
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ham radio magazine

AUGUST 1969

16-foot dish for UHF moon-bounce



HOMEBREW PARABOLIC REFLECTOR

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ham radio magazine

august 1969

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roving editor
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vhf editor
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associate editors
A. Norman Into, Jr., W1CCZ
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art director
Jean Frey

publisher
T. H. Tenney, Jr. W1NLB

offices

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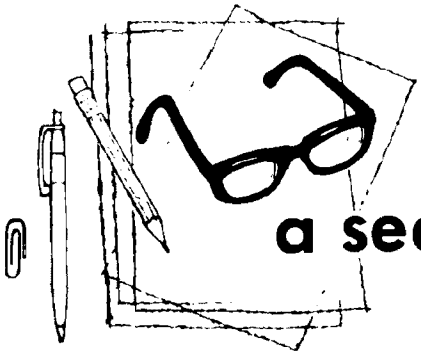
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a second look

by jim fisk

While I was researching some textbook material for an extensive article on vhf fm for the September issue of **ham radio**, I discovered a very curious fact: the standard textbooks did not agree on the differences between phase and frequency modulation. The textbooks that provided a mathematical treatment on the subject tended to be correct, while those that discussed it on a less technical level were usually in error.

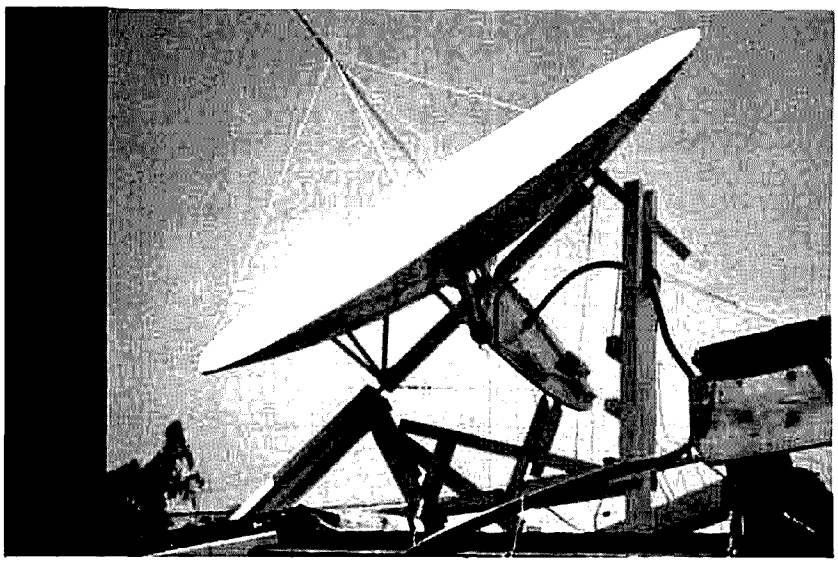
There is a very simple difference between phase and frequency modulation: the frequency deviation of a frequency modulated wave is proportional to the peak **amplitude** of the modulating signal, and **independent** of the modulating frequency: frequency deviation in a phase-modulated wave is proportional to **both** the amplitude and frequency of the modulating signal. The difference is quite subtle, and since both phase and frequency modulation are forms of angular modulation, they are not essentially different, and one can be converted into the other.

This conversion from one to the other is important because although fm offers superior performance in terms of signal-to-noise ratios, phase modulators are more convenient to use with crystal-controlled transmitters. Since fm phase-modulated waves have greater deviation at the higher modulating signal frequencies than pure fm waves, an RC network that reduces the frequency response of the audio system on the higher modulating frequencies takes care of the conversion from pm to fm.

The important thing here is not this particular error, but the fact that errors do occasionally creep into textbooks. In this case, it was probably a case of the blind leading the blind—an early non-mathematical article on the subject was 180 degrees out of phase with the facts. Unfortunately, this article was used as the basis for several simple textbook discussions on fm; later textbook authors used these erroneous chapters as the basis for their words on the subject and the error was promulgated from one textbook to another, right up to the present day. Electronic-engineering textbooks, as I noted earlier, tend to present the matter correctly, particularly if the author has gone to the trouble to include the math.

When you run into a conflict in technical view, the question inevitably arises, who is **the** authority? Although there are dozens of excellent books on the subject of radio communications, there are two texts that stand out above all others: F. E. Terman's "Electronic and Radio Engineering" (McGraw-Hill, 1955), and Keith Henney's "Radio Engineering Handbook" (McGraw-Hill, 1959). If you don't have a copy of one of these at home, at work, or at your local public library, you should consider buying one volume or the other. Summer is a good time to buy too, since school is out, and used book stores near colleges and universities are often loaded up with books of this kind.

Jim Fisk, W1DTY
editor



a large homebrew parabolic reflector

Complete details for a 16-foot parabolic reflector using honeycomb sandwich construction

Peter Laakmann, WB6IOM, 8001 Airlane Avenue, Los Angeles, California

Constructing a large parabolic reflector with amateur means and budgets is by no means an easy job and requires hundreds of hours of effort before the dish is finished, installed, and ready to operate.

However, any serious uhf enthusiast will find out sooner or later that parasitic arrays will not do the job for EME work above 400 MHz. Commercial dishes which will perform as well as my homemade one are extremely expensive ($5000 to $10,000) so they are out of the average amateur's reach. The surplus market isn't much help either—supplies of large surplus parabolic reflectors are extremely limited.

For most uhf and microwave moonbounce enthusiasts the path to adequate system performance will include the construction of a large parabolic reflector. Although many amateurs have travelled this road successfully, even more have failed. Most failures apparently occurred because the builder underestimated rigidity and accuracy requirements for reflectors for the higher frequencies. Other failures occurred because operational pointing accuracy requirements were underestimated.

Errors in homemade reflectors are usually due to gross distortions that extend over a large portion of the dish surface (many wavelengths); these errors have the most effect on gain loss. To obtain full gain (within 1 dB of theoretical) surface errors must be less than 1/32 wavelength rms. At 1296 MHz this is a little more than one-quarter inch. There is no point in building a larger dish with less accuracy because it compounds the construction difficulties without producing any more gain than a smaller, accurately constructed dish.

design

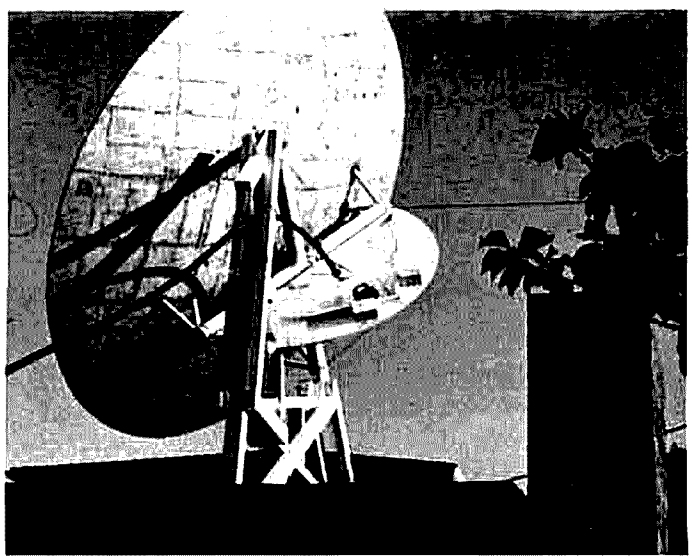
As dish size increases, aiming must be more accurate. A 10-foot dish for 1296-MHz moonbounce must be pointed with an accuracy of better than 1 degree or you will degrade moon echos more than 1 dB. On 1296-MHz I doubt that it is within amateur means to point a 30-foot dish accurately enough to take advantage of its full gain (20 minutes of error). These difficulties become even greater at 2300 MHz and higher.

I settled for a 16-foot parabolic for three reasons: it provides adequate system gain at 1296 and 2300 MHz, it can still be pointed accurately enough, and most of all, it is the largest size that could be built in my garage!

The f/d (focal length to diameter) ratio was chosen as 0.525 for a focal length of 8.4 feet. This is close to the optimum f/d for gain using simple feeds (0.40 to 0.50). The larger focal length allows later expansion to 20 feet and provides more latitude in pointing the dish over a small angle by moving the feed.

The dish is mounted on a polar mount with a fixed declination of plus 12 degrees. The feed is moveable over a range of plus and minus 16 degrees to cover all positive moon declinations. Gain losses due to feed steering are quite small.

Completed dish and polar mount. Coaxial feedline is 1-5/8" Heliax.



materials

The dish is constructed of aluminum honeycomb sandwich. This yields a solid surface with approximately twice as much wind resistance as a typical mesh surface. This is a disadvantage in high-wind areas but is not a factor in Los Angeles where the maximum recorded wind velocity has never been over 50 mph. City building code regulations call for the design of surfaces and support structures to withstand 15 pounds per square foot. At the "parking" angle of 22 degrees—the worst case—this only amounts to 1150 pounds loading.

All surface and attachment joints are made with epoxy. No bolts or welds were used anywhere on the dish or its rear frame structure. However, joints between the frame structure and the dish use fiberglass matting to assure better bonding in these high stress areas.

After the dish was placed on the gimbal on the roof of the house, 1/4-inch bolts were added to secure the rear frame to both the front and rear surfaces of the dish. These were epoxied in also. These bolts are probably not necessary, but I didn't want to find out during a windstorm. Epoxy joints are a tricky matter at best, particularly with aluminum; some turn out very well and some not so well—the difference is usually a result of cleanliness in construction.

One characteristic that makes honeycomb structures attractive for large reflectors is their fantastic rigidity relative to their weight. The dish and frame weighs approximately 230 pounds total. The frame is only attached to the dish at 5 points, yet applying 50 pounds force to the rim of the dish causes

distortions of less than 1/4 inch.

Of the total weight, about 30 to 40 percent is nonstructural and due to the epoxy I used. More careful techniques of applying epoxy without waste could cut dish weight to less than 200 pounds. Dish weight is quite important because distortion due to gravity is usually more significant than windloading since gravity acts all of the time.

This article will dwell in detail on the construction of the dish using readily available materials. The dish is self-supporting when a **rigid** frame is attached to the center and four points about 5 feet from the center. The attachment area must cover at least 0.5 square foot for each point with fiberglass matting to provide uniform loading against the honeycomb sandwich.

I won't go into complete constructional details of my rear frame since I used materials that aren't easily obtained. It is made from four missile containers and honeycomb tubing that was originally intended to be a dissipative shock absorber for shock testing. The four cylindrical missile containers are epoxied together to form an X. A well thought out wood frame should work just as well. The missile containers are essentially 6-inch diameter fiberglass tubing sections with wall thickness of 0.02 to 0.03 inch and reinforced ends about 1/8-inch thick. The photograph on page 14 will clarify the attachment technique.

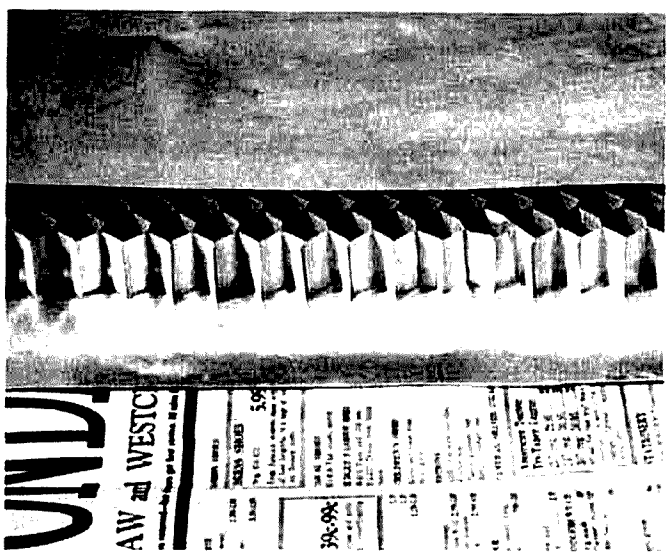
The materials used for the dish itself are fairly inexpensive; total cost is about $250. Considering the performance of this reflector it is a good investment. Other costs are associated with the pattern on which the dish is constructed, chemicals, containers for cleaning and etching, countless pairs of rubber gloves, brick weights, rubber mats, and of course, the frame and gimbal for the finished dish.

A significant factor in the total cost is the amount of extra material which must be purchased when making only one dish. This must be done so sufficient material is always available to complete the job. Also, the honeycomb material and epoxy cannot be bought to the exact quantities required as it is shipped in standard increments. If a group of amateurs gets together to make several of these dishes as a team project, a lot of money and tooling could be saved.

honeycomb construction

A simple honeycomb sandwich consists of two faceplates and a core that looks like a bee's honeycomb. The core is joined to the faceplates by bonding or brazing. This constitutes a rigid sandwich that does not bend easily. Before bonding, the honeycomb core can be deformed to shape it to a mildly curved surface. When bending the honeycomb in one plane it naturally takes on a concave/convex "saddle" shape and resists the doubly concave or convex shape used in parabolic reflectors. With evenly applied pressure it will take on the desired shape if the curvature is mild, but when deformed beyond its limits it collapses.

Honeycomb sandwich used in the construction of the 16-foot dish.



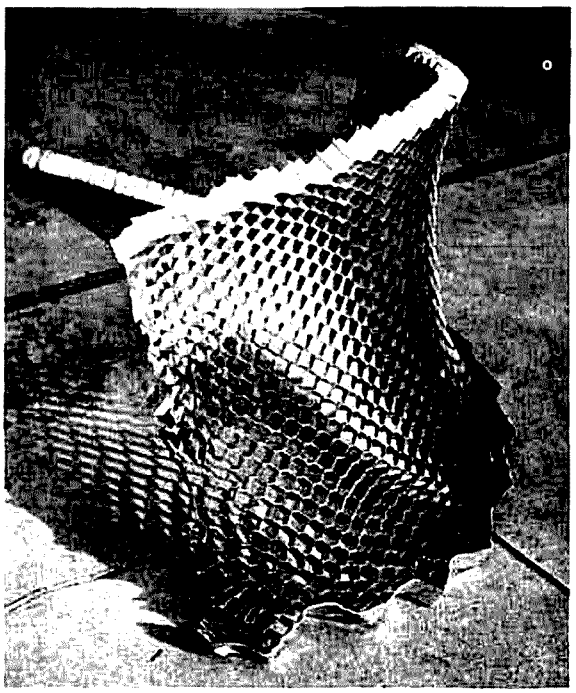
The material used for the two faceplates does not have to be identical to the core material. The honeycomb core is available in paper, aluminum, epoxy fiberglass and steel. Rigidity is improved by increasing either faceplate thickness or cell height; rigidity is not affected by cell size or thickness of the core material. Cell size and core thickness affect compressive strength.

Honeycomb sandwiches can be tailored to the exact rigidity and strength require-

ments you need without overdesigning one to get the other. However, certain combinations of cell height and faceplate thickness result in a minimum-weight design for a stated rigidity. These combinations can't always be used, though, because the faceplate skin becomes excessively thin and the core is too high to bend very much.

The aluminum honeycomb I used for this dish has a cell size of 3/4 inch, a cell height of one inch and a wall thickness of 0.003 inches. The faceplates are 0.008-inch aluminum. Believe it or not, the compressive strength of this sandwich is 50 pounds per square inch!

Honeycomb material takes on a saddle shape when deformed.



Core cost is about $.35 a square foot. Total weight of the core material used in the dish is 30 pounds. The minimum weight sandwich would have consisted of honeycomb 2 inches high with 0.002-inch faceplates; the result would be 14 pounds lighter than my chosen design. However, 0.002-inch faceplates are impractical and too easily damaged. In addition, the 2-inch honeycomb would not conform to the dish curvature. The 8-mil faceplate material I used is quite sturdy, and even permits walking on the surface without damage.

Bonding the honeycomb core to the faceplates is a simple process. The two pieces of sheet metal are coated with a thin coat of epoxy (about 10 mils) and the core is sandwiched in between. Uniform pressure is then applied, and the epoxy is allowed to set. The epoxy creeps up the walls by capilliary action and forms a fillet that is higher than the coating thickness.

the epoxy

Choice of epoxy for honeycomb bonding involves many factors: it must be a room-temperature curing type, it must have several hours of pot life, and it must be thicksotropic* to keep an adequate film on the sheets to be bonded. But above all it must be capable of achieving a structural bond to aluminum. Aluminum is probably the most difficult material to bond. The epoxy I used is the same as that used in the honeycomb construction[2] of the 60-foot reflector at Massachusetts Institute of Technology's Lincoln Laboratory and it does an excellent job of bonding aluminum.

Most of the other epoxies that I am familiar with do not achieve a true bond to aluminum; the aluminum can usually be peeled off the joint with very little force, leaving no trace on the aluminum of more than a superficial joint. The epoxy I used is General Adhesive and Chemical Company's type 88-110. This material apparently attacks the aluminum to create a structural bond. The force required to peel off the aluminum almost tears the metal before the epoxy gives. I strongly recommend that you not make any substitutions; also use the epoxy strictly in accordance with the manufacturer's instructions. Don't use any additives to prolong the pot life; I tried this with disastrous effects on bond strength.

The epoxy is quite reactive. Skin contact should be avoided as it can cause allergies in some people (generally true of all epoxies). Use rubber gloves all the time and

* Thicksotropic. Denotes a thick consistency that does not easily run off inclined surfaces. It is not quite the same as high viscosity as that term describes a phenomenon related to friction due to movement.

work only in well-ventilated areas.

The recommended ratio of resin to hardener is 7:1 by weight. It isn't too critical, however, so an ordinary household scale will weigh it accurately enough. In general, adding more hardener decreases both pot life and viscosity; neither effect is very desirable. However, when working with the material on cold days the ratio can be changed to 5:1 to maintain a viscosity that can be more easily worked. Pot life is increased by lower temperatures so two hours of pot life are still available.

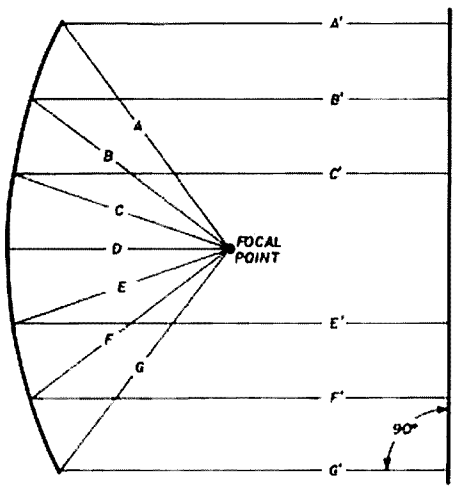
The resin is sold in 5-gallon cans. Depending on how carefully it is used, anywhere from 5 to 7 gallons are required for the complete parabolic reflector. I recommend that you get an extra can of hardener since it seems to get used faster than the resin; the hardener is shipped in one gallon cans. Cost of the epoxy is $9.62 per gallon.

the pattern

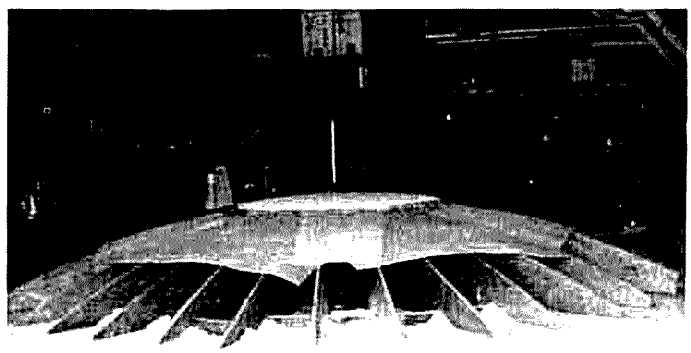
The first step in the construction of any parabolic reflector is the creation of a pattern that is used as a reference surface. I made my pattern out of plywood, window screen and concrete as these are inexpensive. The pattern is hollow and the concrete is about one inch thick. About 5000 pounds of concrete are required (about 1 cubic yard), so it is best to buy it already mixed. I didn't realize how much was required and bought it by the bag. This resulted in a lot of work and was actually much more expensive than it would have been if delivered ready-mixed from a truck.

The first step is to clear out a large flat area (I used my garage). Cast a 1½-inch diameter pipe in concrete in the center of the area. This is the minimum thickness pipe that can be used as the center of rotation for a parabolic ruler that is to define the parabolic surface. Thinner pipes bend too much. The pipe should be filled with concrete to stiffen it even more.

fig. 1. This sketch shows how you can generate a parabolic curve using a piece of string. In this drawing, A + A' = B + B' = C + C', etc.



Partially finished pattern.



Next make a parabolic radius ruler from ¾-inch thick plywood. The curve can be generated numerically or by means of a string, using the definition that a parabola is equidistant from the focal point and a straight line (called the directrix) that lays behind the parabolic curve.

Attach the radius ruler to two bearings at least two feet apart; make sure the bearings fit the center pipe tightly. Also, make absolutely sure that the bearings are perpendicular to the radius ruler. When the ruler is attached to the center pipe, it should be stiff enough so it doesn't deflect more than ⅛ inch when a 10-pound force is applied to the end of it.

By rotating the radius ruler 180° around the pipe and taking height measurements against a true horizontal line, any errors from a non-parabolic curve can be found and corrected before further work is done. It's all right if the axis of rotation (the center pipe) is not

quite vertical, but it **must** be perpendicular to a tangent drawn to the parabolic curve in the center.

Next cut 32 plywood radius sections approximately parabolic in shape but about 1/2 inch less than the height to the radius ruler from the floor. Sixteen of these sections should be cut for the full radius of the dish; the other sixteen should be only half as long and are placed around the rim of the pattern as shown in the photo.

The 1/2-inch height difference between the plywood pattern and the parabolic ruler is adequate since the weight of the concrete causes some sagging between pattern sections and results in a mean concrete thickness of about 1 inch.

Anchor the plywood sections together with nails and short sections of 2x4's. Starting at the center, run steel wire around the pattern every 3 inches of radius and tighten by wrapping it around nails in the plywood. Cover the pattern with window screen and nail it to the plywood. Finish only about half of the total radius at this time so the center of the pattern is still accessible.

Starting from the center toward the outside, put wet concrete on the window screen and smooth it out with the radius ruler. Use gravel filled concrete for the initial pass and fine concrete for the top surface. Be careful not to produce excessive wear on the ruler by "grinding" with it. Stop when about one-third of the total radius is smooth; repeat the process on following weekends for the rest.

Add window screen and steel wire as you expand the concrete covered area toward the rim. After the concrete surface is finished, use plaster of Paris or similar materials to smooth out the concrete by hand and obtain a shiny surface without a sandy appearance. Remove the parabolic radius ruler. Now paint the pattern so it is dust free and smooth. It is now ready to be used.

etching the aluminum

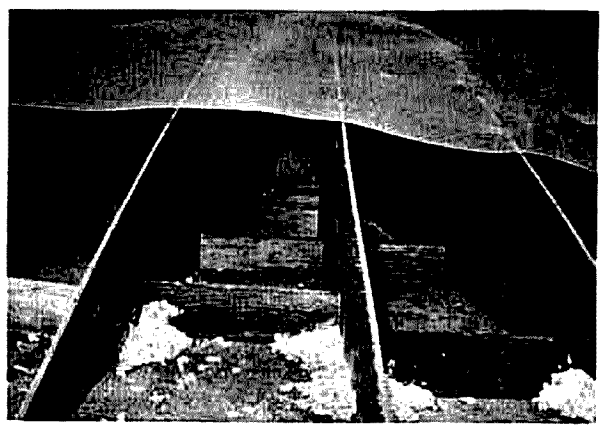
The aluminum used for the faceplates of the sandwich must be chemically etched to assure good quality bonds. The aluminum honeycomb should be etched also; however, the material delivered to me by Hexcel was found to be adequate for bonding without an etching process. The epoxy bond to the honeycomb is naturally stronger anyway because of larger area and edge roughness of the honeycomb. It's a good idea to make a couple of trial bonds with your material to check for any problem in this area.

The 1x2-foot aluminum sheets are etched in a chromic acid solution at a temperature of approximately 120° to 150° F. This size represents the largest faceplate that can be made to conform to the curvature of this reflector. If you plan to make a different size dish, or different f/d ratio, it would be a good idea to experiment with the size of the sheet on the pattern to determine the best size. The 1x2-foot size is not too large to be etched in a 32-gallon polyethylene trash can; in fact, up to three sheets can be etched at

Friend Mike Mikalides smears concrete on the pattern while WB6IOM handles the ruler.



Closeup of the pattern showing plywood ribs and window screen.



one time. Two trash cans, one inside the other, are required for safety and rigidity.

Fill the trash can about half full with regular tap water and slowly add the sulfuric acid. Then dissolve the sodium dichromate. The heater for the solution is made from number 18 teflon hookup wire. The length of the wire is adjusted to absorb about 300 watts from a low-voltage filament transformer. Time to reach temperature equilibrium is about 8 hours, so power should be applied the evening before for use the next morning.

The chromic acid etching solution is not too dangerous but rubber gloves should be used while working with it. Safety goggles are a good idea too and at least regular eye glasses are required. The solution is about as potent as battery acid but it is poisonous due to the chromate content. Do not do any etching in an enclosed area.

The aluminum sheets should be immersed for about three minutes each, or until there is clear evidence of etching. There is some bubbling action but it is much milder than a caustic etch. Don't substitute a caustic etching process for the chromic acid; it does not leave a clean surface.

After etching, the aluminum sheets should be placed in a small children's swimming pool that is supplied with a continuous flow of tap water. After rinsing for a few minutes the sheets can be removed and hung up in a vertical position to dry. After drying the sheets may be stored for later use, but don't get any grease or fingerprints on them.

bonding the faceplates to the core

Bonding between the three components is done on the plywood-concrete pattern. One side is bonded to the honeycomb core at each session. The top faceplate is bonded in place only after the bottom faceplate has had time to cure (about two days at 70° F. for full hardness). The maximum number of faceplates that can be worked in one session by a single worker is 8. It's a good idea to start out with fewer faceplates at first to gain experience so that the epoxy does not start to cure in the middle of a bonding session. The pot life of the epoxy is about two hours. Bonding 8 plates requires about 2 pounds of epoxy.

The following explains the necessary procedures in detail. I highly recommend that this procedure be followed religiously as I cannot possibly cover all the possible pitfalls that might occur with another approach. The dish should be built up from the center toward the rim, and then expanded in both circumferential directions until the ends meet.

1. Place newspaper on pattern surface and tape together with masking tape. Tape ends of newspaper to pattern. Newspaper area must be slightly larger than area to be worked that day. The newspaper will act as a "mold release" so that the finished dish does not stick to the pattern.

Completed installation shows rear frame attachment points. This photo was taken before the feed was installed.



2. Run masking tape down one side of the first faceplate sheet. Use tape at least one inch wide and center the edge of the sheet on the masking tape. Apply a thin coating of epoxy along the edge of the aluminum sheet (about 2 inches wide) with a putty knife. The epoxy is applied to the side that is to be face up; the masking tape is applied to the side facing the pattern. With experience you

can judge the thickness of the coating by the deep red color of the epoxy.

3. Take the second faceplate sheet and press it against the masking tape, allowing for a sheet overlap of approximately 1 inch. Continue to add faceplate sheets in this way until the area to be worked that day is finished.

4. Tape the aluminum sheets down against newspaper and the pattern in a few spots to prevent them from sliding.

5. Apply epoxy to the entire area.

6. Cut a piece of honeycomb that will leave the outer 2 inches of the aluminum sheet free. The honeycomb can be cut with ordinary household scissors.

7. Press the honeycomb against the aluminum sheet and move it back and forth about ¼ inch to make sure the edge is well covered with epoxy.

8. Place bricks around the outside edge of honeycomb to apply pressure uniformly around the edge. This should result in compression throughout the area. If not, add bricks as required.

9. Remove all epoxy from the sheet metal not covered by honeycomb core (use a putty knife). A clean edge is required to facilitate bonding of the adjacent area. A particularly troublesome area is where the aluminum sheets overlap. After cleaning this area, place a small piece of masking tape over it to keep the sheets in compression.

10. Inspect the honeycomb fillets to make sure that the honeycomb is bonded throughout. The center is sometimes troublesome, but the edges cause the most problems. If the honeycomb is not in compression here it makes bonding of adjacent areas very difficult as there is a discontinuity in height of the sandwich. If any areas are found that do not have sufficient epoxy it can be added at this time.

11. Let the epoxy cure. Tools and hands can easily be cleaned with alcohol. However, fully cured epoxy cannot be dissolved with alcohol.

12. (At least two days later.) Remove bricks. Coat the appropriate number of faceplate sheets with epoxy and press them onto the honeycomb. Tape together with masking tape. Overlap between sheets should be approximately 1 inch. Leave approximately one inch of honeycomb around the periphery of the sandwich uncovered.

13. Tape the sandwich down to pattern to avoid sliding and cover it with newspaper. Place rubber mats over the entire honeycomb area and cover entire area with bricks (small ends down to get the most pressure per unit area). If necessary, tape the bricks down to avoid sliding.

14. (At least two days later.) Remove the rubber mats and newspaper and proceed to an adjacent area. Remove all masking tape under the new area to be worked and start at step **1.**

table 1. Materials and tools required for constucting a 16-foot reflector. Construction materials include ample reserve for normal waste. Pattern material and rear frame not included.

construction materials

300 square feet 3/4-inch ACG aluminum honeycomb core, 1-inch high; Hexcel Products, Inc., Havre de Grace, Maryland

500 square feet .008-inch aluminum sheet, 1100H18 alloy cut into 1x2-foot pieces

10 gallons 88-110 epoxy, part A and B; General Adhesive and Chemical Company, Nashville, Tennessee

10 square feet of fiberglass mat for rear frame attachment

tools and disposable materials

5 gallons sulfuric acid

10 pounds sodium dichromate

100 standard 5 pound bricks

8 rubber mats, ¼-inch thick, each 2 square feet

2 polyethylene trash cans, 32-gallon

1 small plastic children's swimming pool

denatured alcohol for cleaning tools and hands

scale for weighing epoxy (will be ruined sooner or later)

large metal shears

quantities of old newspapers

low voltage transformer (5 to 10 V), 300 W or more

100 feet number 18 teflon insulated wire

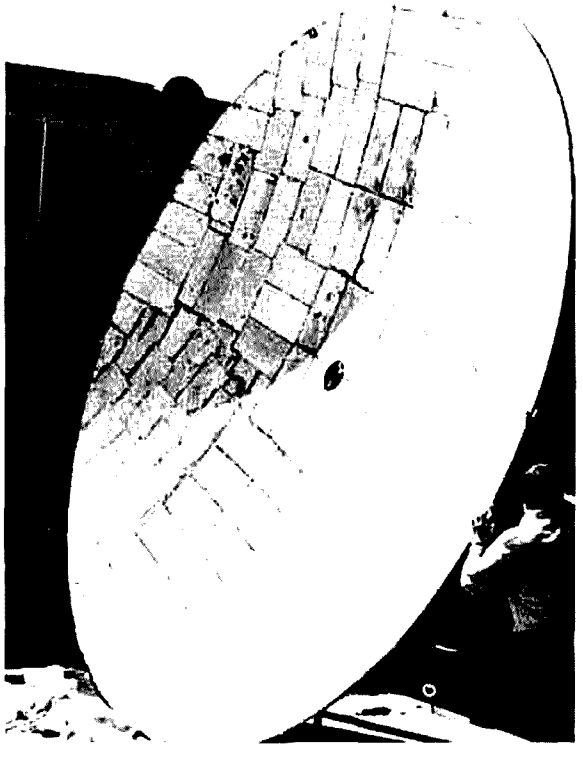
rubber gloves, masking tape and putty knives

When placing the top sheet of the sandwich over adjacent honeycomb areas (areas (areas that were worked in two different days) sometimes the sheet will not lay flat because of height differences (see step 10 above). If this happens, the higher core material can be cut down until the faceplate sheet lies flat.

Adjacent honeycomb cores should be meshed as well as possible. There should never be a gap between sections. If a gap occurs accidentally, it can be filled by a small piece of honeycomb liberally coated with epoxy. The small piece can be compressed while cutting it to the height of adjacent sections. The honeycomb core sections should be placed in an irregular fashion on the pattern so that gaps between sections do not run the full length of the dish.

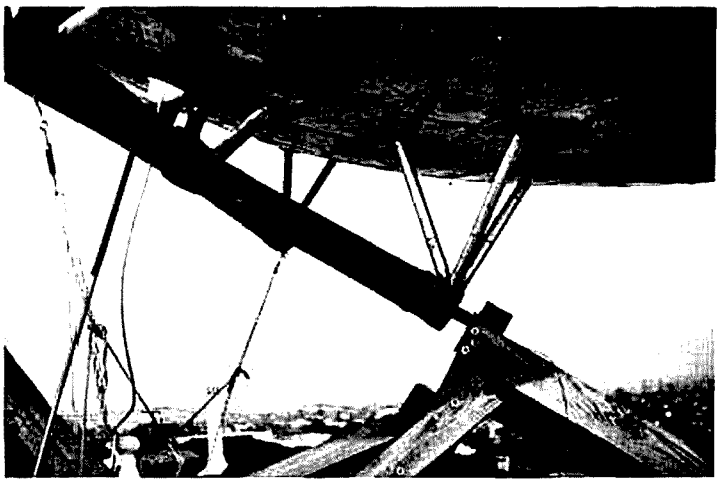
I cannot overemphasize the importance of using as little epoxy as possible to bond the faceplates. The weight of the epoxy is the largest single contributor to the final weight of the structure. The 10-mil coating applied to the faceplates weighs 60 pounds.

Just after removing the 230-pound reflector from the pattern—now to put it on the roof!



Any other unusual additions of epoxy to cover gaps between honeycomb areas, to finish the rim of the dish, etc., add to this.

The epoxy contributes very little additional rigidity to the structure and by its weight does more harm than good. It is also expensive. A coating as low as 5 mils is sufficient for bonding but it is difficult to achieve a sufficiently uniform coating at that thickness. I suggest you keep a running estimate of the total amount of epoxy used vs dish area covered to maintain control over dish weight and the epoxy necessary to finish the project.



Closeup of the x-shaped rear frame and lower gimbal bearing.

the rear frame

After the reflecting surface is finished the rear frame should be attached while the dish is still on the pattern. This defines the position of the attachment points with respect to the backup frame.

The frame should be rigid with respect to the forces to be expected from gravity and wind loading. For a 16-foot reflector, a 50-pound force applied to any end of the four-point frame should not cause more than 1/16-inch movement. The frame I used weighs about 40 pounds and meets the rigidity requirements.

Bearings made from standard plumbing components are attached to the ends of the fiberglass tube. On the reflector end these

consist of one-inch pipe flanges screwed into a wood plug that is epoxied to the fiberglass tube. On the gimbal end two-inch pipe flanges are used; six-inch long pipe nipples are screwed into these flanges. These nipples rest on the one-inch pipe flanges on the reflector frame. For safety, a two-inch-long one-inch nipple is screwed into each of the one-inch pipe flanges on the frame. This makes a crude but adequate bearing for the loads encountered.

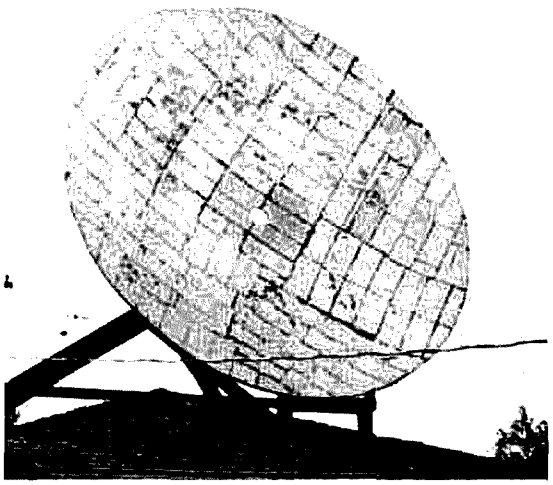
The points of the dish used for attaching the frame should be well cleaned with coarse emery cloth just prior to bonding. To distribute the windloading forces, a minimum of 0.5 square feet is required for each of the attachment points. Each of the attachment points may be broken up into three smaller areas—I did this to use the two-inch diameter honeycomb tubing I had available.

Use a putty knife to apply a thin coat of epoxy to the frame and the dish surface. Then apply a thick layer of epoxy-coated fiberglass matting between the dish and the frame member and keep it under compression until the epoxy cures. You may want to add more epoxy to the top of the fiberglass (surrounding the frame member) to increase its stiffness.

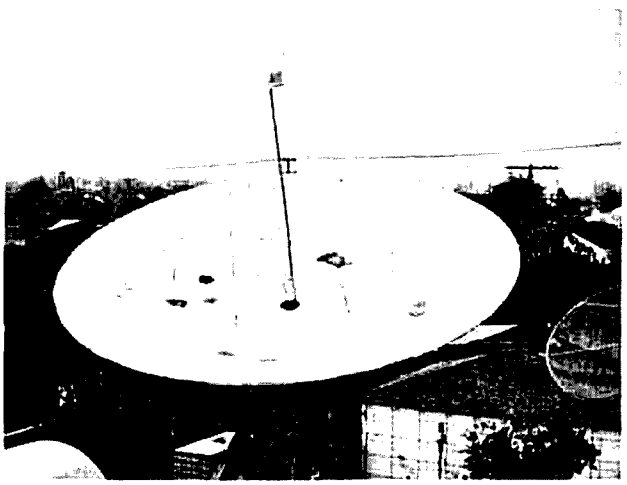
I didn't attach the rear frame to the exact center of the dish because I wanted to maintain access to the center hole for the steerable feed arrangement.

removing the dish from the pattern

Four or five workers are required to lift the dish off the pattern. Pick it up at the rim. After removing it from the pattern you can remove all the newspaper still attached to the bottom. The dish is light enough to be placed onto the roof without the help of a crane. The high end of my gimbal is 23 feet above the ground but five people were able to mount it under winds of approximately 15 mph.



Homebrew 16-foot parabolic reflector sits in a polar mount on the garage roof.



Newly installed reflector with a temporary feed for sun noise measurements.

The dish should be painted a lusterless white to partially eliminate reflecting the sun into the feed. Since this dish has a solid shiny surface it could produce undesirable heating of the feeds in the daytime. Also, it could be hazardous working around the focal plane while making sun noise measurements.

Accuracy of the reflector appears to be better than $\frac{1}{8}$-inch rms. This is good enough for full gain use up to 3300 MHz, and the dish should still have considerable gain at X-band.

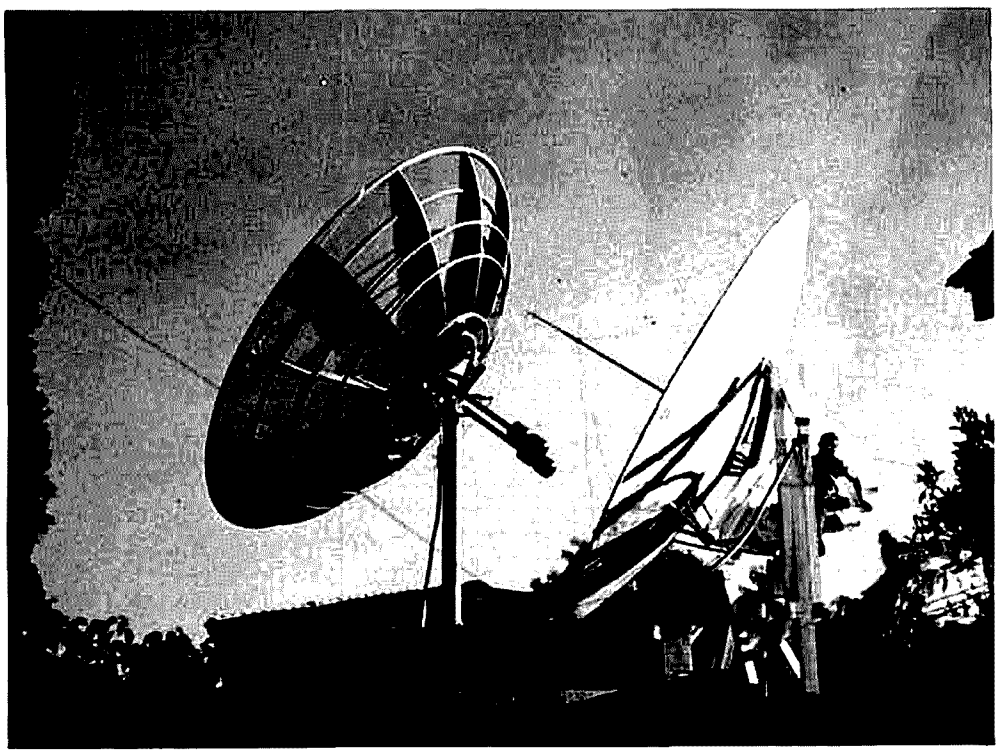
the mount

The drive and counterbalance disk I use with the 16-foot reflector can be seen in the photographs of the completed system. The

disk is in the form of a half circle with a radius of 50 inches, and is made from 3/4-inch exterior grade plywood. The edge of the disk is lined with a deep metal channel, 3/4-inch wide. This channel guides a half-inch nylon rope (available at boating stores) around the circumference of the counterbalance disk. The nylon rope also runs once around a half-inch bolt—between two washers. When the washers are clamped together with a nut, the dish is locked into position. When the dish is "parked," it is held by a safety chain since the nylon is not intended to lock the dish in high winds.

The drive disk is attached to one of the cross members of the reflector's rear frame (see photo) and is held in the polar plane with guy wires. The edge of the drive disk is marked with hour-angle calibrations. The large circumference of the drive disk results in almost one inch per degree for these important calibration points. As I pointed out earlier, large reflectors must be pointed very accurately for maximum performance, so widely spaced calibration points are very helpful.

Counterbalance for the reflector is provided by nine common red bricks. Six of these bricks are strapped to the center of the drive disk and three are mounted at the lower end of the fiberglass tube. These weights provide balance for all positions of the large reflector.



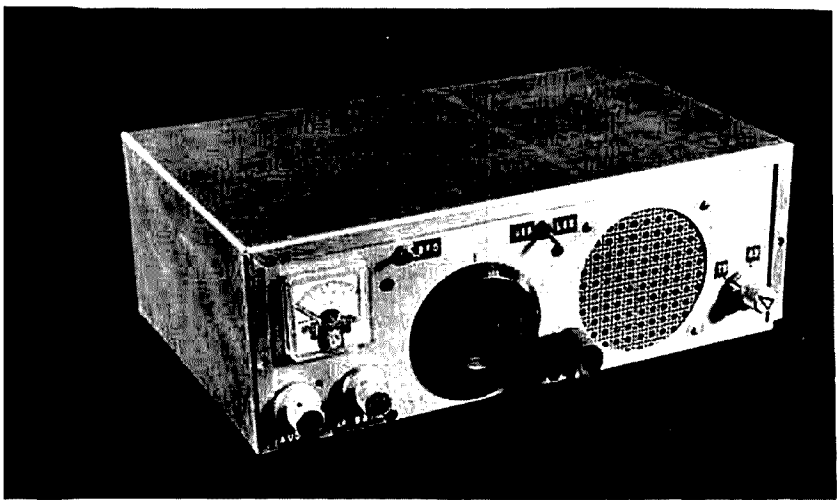
Polar-mounted 16-foot parabolic reflector is to the right. The commercial 10-foot dish in the az-el mount to the left has been used for successful two-way moonbounce contacts with G3LTF and W2NFA.

All the structural members of the polar mount I used are either 6x6- or 4x6-inch Douglas fir timbers. Roof loading is minimal since the vertical members of the mount go through the roof to the floor of the garage.

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ham radio



solid-state Q5er

Replacing hot tubes with cool transistors makes this twenty-year-old veteran better than ever

Robert J. Grabowski, W5TKP, New Caney, Texas 77357

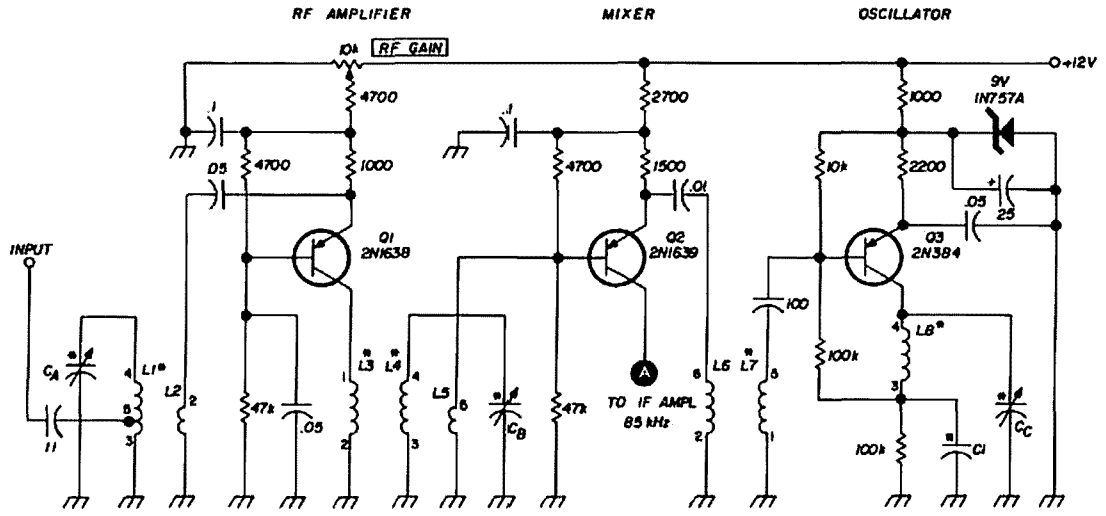
Large quantities of military surplus radio equipment became available shortly after World War II, and amateurs have made extensive use of the many high-quality pieces. Some of the equipment was used without conversion or modification, some was used as a source of excellent precision components, and some was altered for use in the amateur bands. One of the most popular pieces of military surplus is the low-frequency receiver designated BC-453 and commonly called the Q5er by many hams. Many articles on conversion and use of this receiver have appeared in radio magazines.

When the command receivers first appeared on the surplus market, the value of the BC-453 was not recognized. Some dealers offered this model as a free bonus with the purchase of one of the others in a set. However, when it was shown that it could be easily added to an amateur communica-

lions receiver to greatly improve selectivity, the popularity and price of the BC-453 immediately increased. This use was responsible for the origin of the term Q5er. The history of the command set series has been reviewed by White;[1] a typical schematic and coil data are included in the review.

conversion to solid state

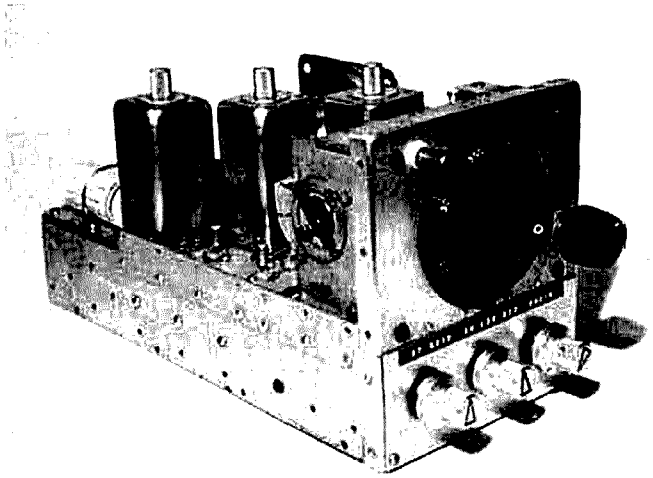
About five years ago I considered replacing the tubes in the BC-453 with transistors. An experimental model was converted with excellent results. The total power consumption was less than that required to operate the heater of one of the tubes in the original



C1	690 pF total including 20 pF variable	L5	15 turns no. 30 next to L4
L2	10 turns no. 30 next to L1	L6	5 turns no. 30 next to L7

fig. 1. The rf amplifier and frequency-converter stages used in the solid-state Q5er. The biggest modification task was adding the coupling coils to provide a correct impedance match to the transistors. Parts marked with an asterisk are original BC-453 components.

This solid-state Q5er draws less current than one of the tubes in the original version.



model. The conversion circuit was the subject of a MARS tech-net program in December 1964. The initial model involved altering the rf and i-f coils by adding windings to obtain a favorable match for the transistor low-impedance input. Although not considered difficult, modification of the coils is not a task to be undertaken by anyone without radio construction experience.

A second unit was converted, and on this model the rf and i-f coil locations on the original chassis were not changed. Circuit diagrams of this model are shown in **figs. 1, 2** and **3.** Minor modifications were easily made to the rf coil assembly, and a capacitive divider was added to each of the i-f cans to provide an impedance match for the i-f transistors and the detector input.

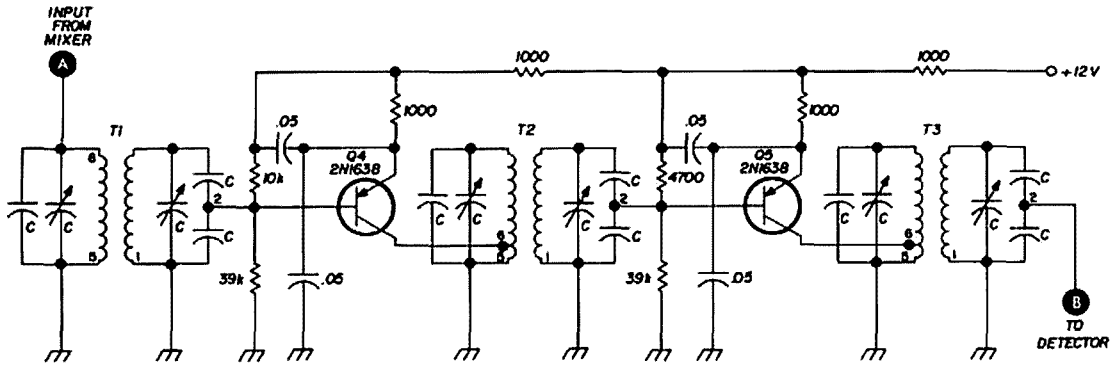


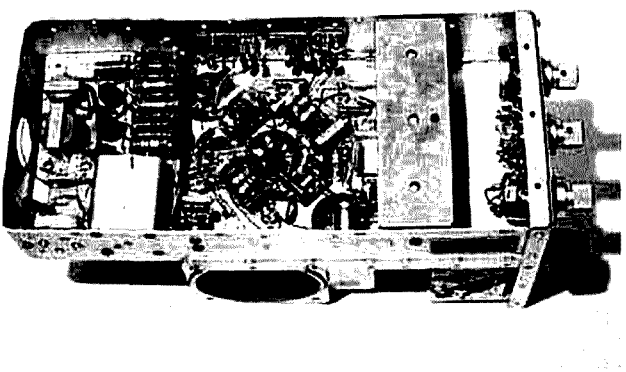
fig. 2. The i-f strip. A capacitor divider was added to the transformer secondaries (details and values for "C" are given in fig. 4.) T1, the first i-f can is coded red; T2, yellow and T3, blue.

rf, mixer and oscillator

The rf-amplifier transistor is used in a grounded-base circuit (**fig. 1**), because neutralization is usually not required, and moderate gain is available. A ten-turn coil, wound on the form and positioned close to the original input coil, provides the low-impedance source to drive the emitter of the input transistor, Q1. Another small coil (15 turns) added to the rf amplifier coil provides drive to the mixer, Q2.

The oscillator transistor is a 2N384. Other high-frequency types may be used, but I found that this transistor produces an excellent sine wave of nearly equal amplitude over the frequency range covered. A five-turn pick-up coil is added to provide drive coupling to the mixer emitter.

Under-chassis view of the Q5er; solid-state audio section is to the rear.



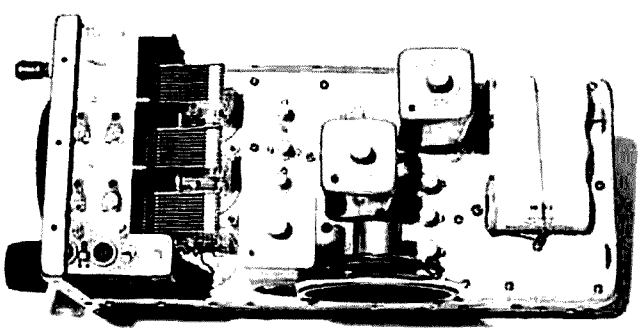
i-f amplifier

The circuit for the i-f amplifier is shown in **fig. 2.** Modification of the i-f transformers is simple, and the same change is made in each unit. Addition of a 20-pF capacitor in parallel with the original (180 pF) and a 0.002-pF disc capacitor in series with the total 200 pF provides a convenient tap for the impedance match and retains the effective 180 pF required to tune each coil to 85 kHz. The modified coil circuit arrangement is shown in **fig. 4.** The three i-f transformers are identical, except that the primary and secondary of the first i-f can are not tapped.

detector and bfo

The detector and bfo circuit is shown in **fig. 3.** The detector is a type suggested by

Top view of the Q5er shows the battery-pack power supply and transistor layout.



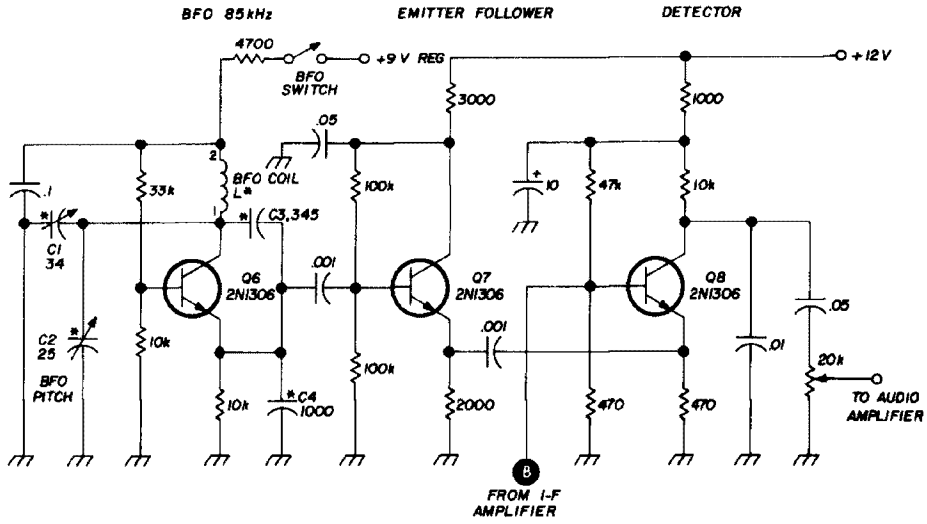


fig. 3. Bfo and detector. Sufficient signal is available to drive a simple two-stage audio amplifier. Parts marked with an asterisk are original BC-453 bfo components.

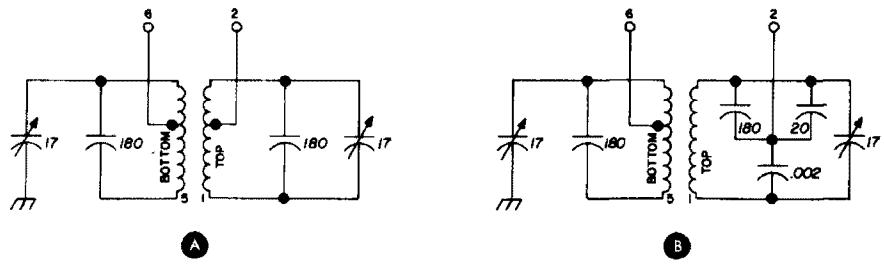


fig. 4. BC-453 i-f transformer detail. Original is shown in A. A series-parallel capacitor arrangement was added in B to provide an impedance match to following stages.

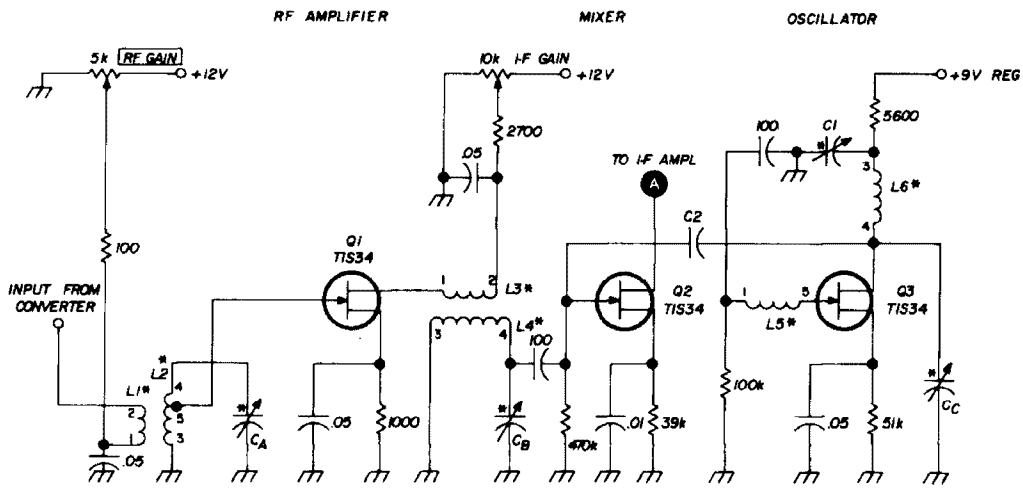
Stoner[2] and works well with a-m, cw and ssb. Detection of cw and ssb signals requires injection of a signal from the bfo.

The audio signal at the detector output is sufficient to drive a simple amplifier. A single voltage-amplifier stage followed by a power amplifier will provide adequate gain.

an fet version

The availability of field effect transistors at prices competitive with bipolar transistors added new interest toward converting the BC-453 to complete solid-state. Since the fet is considered to be a solid-state counterpart of the vacuum tube, the possibility of mak-

fig. 5. Rf, mixer and local oscillator using fets. The BC-453 coils were used without change in this version. Coils L1 through L6 are original BC-453 coils.



C1 690 pF total, including 20 pF variable

C2 gimmick, approximately 1 pF (2 pieces hook-up wire twisted over 1 inch)

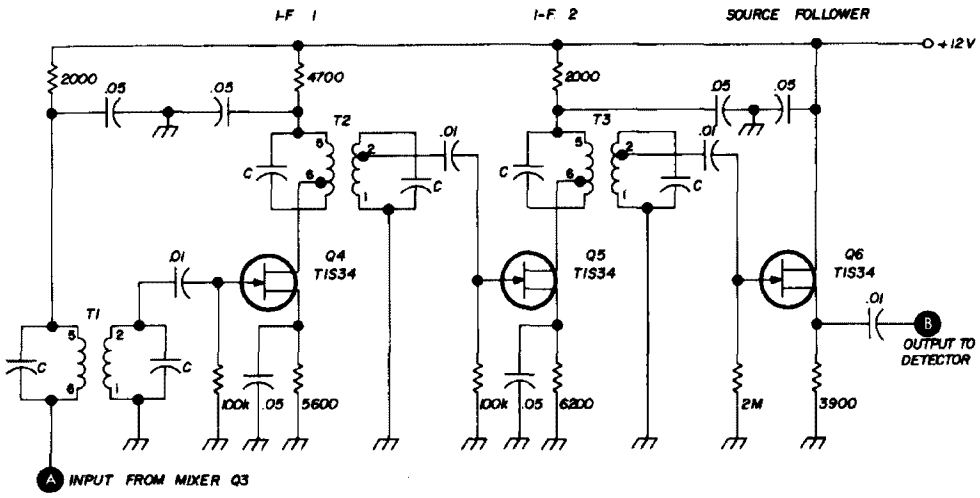


fig. 6. The fet i-f amplifier. T1, T2 and T3 are the original BC-453 85-kHz i-f transformers; capacitors labeled C are the 180-pF trimmers in the i-f cans.

ing the modification without coil alteration was investigated. Again, the conversion was easily made, and results were excellent. The fets are the one-dollar variety (TIS34). The circuits for the fet model are shown in **figs. 5, 6** and **7.** It should be noted that the original coils are used **without** alteration. Future plans include the addition of avc, noise limiter and a Q multiplier.

mosfet converter

Fig. 8 is a schematic of a frequency converter for the 75- and 40-meter bands. This unit uses one of the new mosfets, and more than enough gain is available. Strong signals are easily copied without an antenna (receiver case removed).

Some models of the BC-453 have an rf can containing only one winding, which is connected to the antenna post through an 11-pF coupling capacitor. In this case, the converter mixer drain load may be replaced with an rf choke and the mixer output taken from the drain side (conventional capacitive coupling).

The 3500-kHz crystal may be changed to suit individual needs or availability. A military surplus 3450-kHz crystal is a good substitute. With this crystal, the 75-meter phone band tunes from 350 to 550 kHz, and the

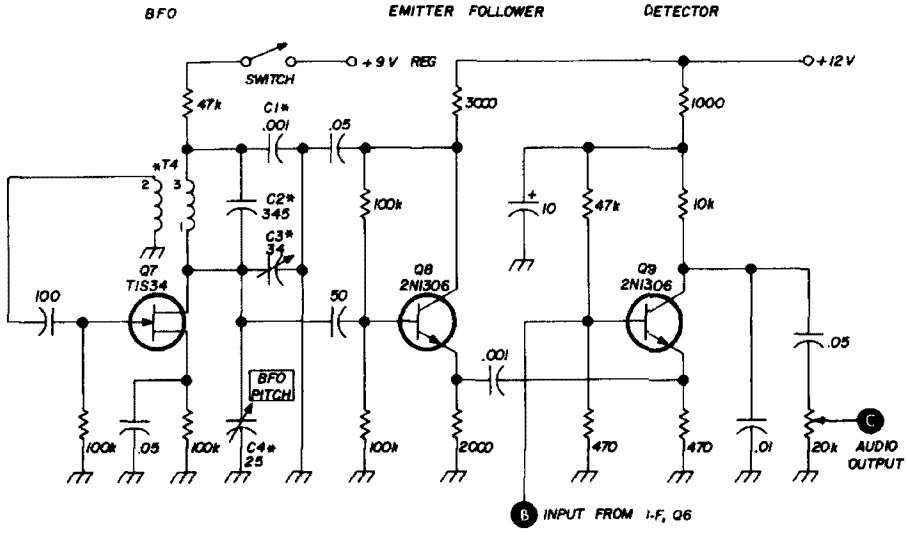
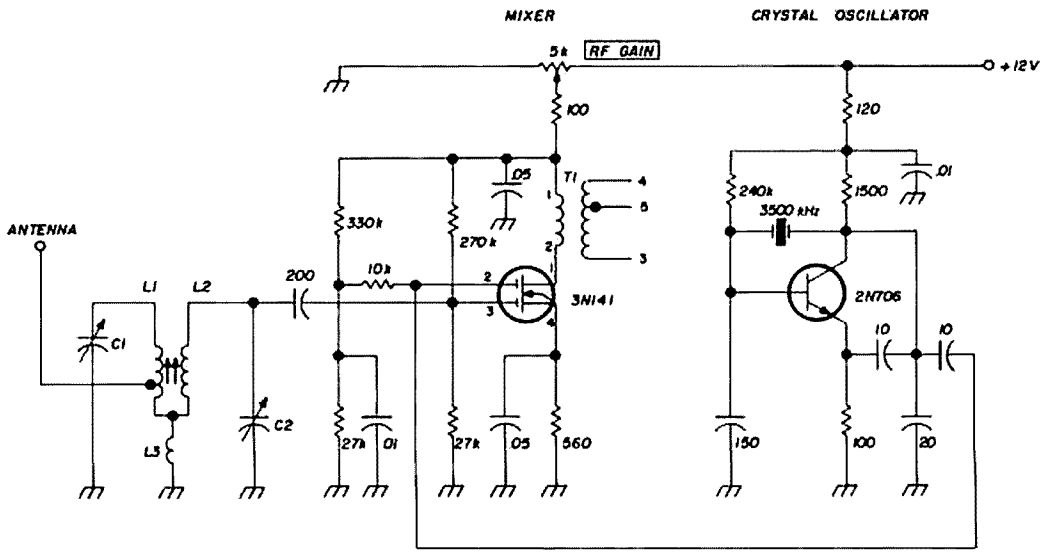


fig. 7. An fet is substituted for the 2N1306 bfo in this version. Remaining stages are identical to those in fig. 3. Original BC-453 parts are marked with an asterisk.



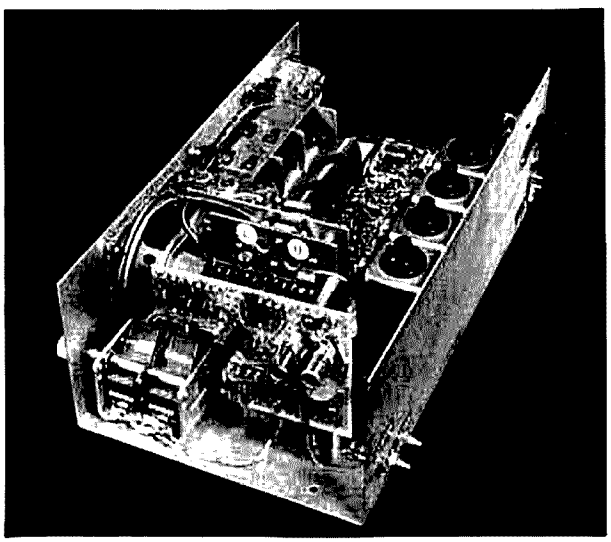
C1, C2	dual 365-pF variable	L3	0.2 μH. 10 turns no. 20 on 1 W resistor
L1, L2	7 μH. 26 turns no. 26 on 3/8" form, spaced 1 inch, antenna tap at 5 turns	T1	BC-453 antenna coil

fig. 8. Mosfet converter for 75 and 40 meters.

40-meter phone band tunes from 300 to 400 kHz. The crystal oscillator circuit produces many usable harmonics and has been used to provide a 144-MHz signal from an 8-MHz crystal for receiver tests.

Band changing is accomplished by merely adjusting the 365-pF tuning capacitor. With the coil values indicated, 75 meters peaks when the setting is approximately at 280 pF, and 40 meters is in tune at a setting of 75 pF. WWV signals at 10 MHz may be copied at 350 kHz on the dial if the capacity is decreased to about 20 pF (3450-kHz crystal in the converter).

This repackaged version of the solid-state Q5er has a built-in 75- and 40-meter converter. A front view of this model is shown an page 20.



caution note

Extra precaution should be used when working with mosfets, since static charges can easily destroy some of the junctions. The units are sold with all leads in contact —usually held by a small eyelet. The mosfets may be handled without extreme caution if a fine strand of bare wire is wrapped around all the leads near the case before removing the eyelet. The wire may be removed after the unit is in the circuit.

I hope these conversion suggestions will stimulate new interest in the BC-453 and perhaps add life to the units considered part of the usual junk box. Use of integrated circuits in a fourth unit is currently in progress.

references

1. Gordon E. White, "Command Set Receivers For All Frequencies—The Easy Way," CQ, January 1967, p. 21.
2. Stoner and Earnshaw, "The Transistor Radio Handbook," *Editors and Engineers, Ltd.*, 1963, Summerland, California.

ham radio

distortion in fm systems

Adjustment of receiver and transmitter for optimum performance in the fm mode

Carl C. Drumeller, W5JJ, 5824 N.W. 58th Street, Oklahoma City, Oklahoma 73122

There has been considerable interest recently in popularizing fm on the two- and six-meter bands. Commercial equipment, obsoleted by FCC rulings requiring narrower frequency deviation, has been adapted by some amateurs. Others have modified their a-m equipment for fm with varying degrees of success.

Adapting a-m transmitters for fm generally consists of frequency modulating a vfo and operating the following stages in class C. For receiving, slope detection is used, or a discriminator is substituted for the a-m detector.

Whether you use modified commercial equipment or adapt your a-m set for fm, there are some fundamental concepts that should be observed to avoid unsatisfactory performance when transmitting and receiving parameters haven't been standardized. The following is a review of causes of nonlinearities in fm systems, using both modified a-m receivers and discriminators. A discussion is also presented on adjusting frequency deviation in fm transmitters.

slope detection

Consider first the use of an a-m receiver for the reception of fm signals. **Fig. 1** illustrates the principles. In the example shown, the receiver is tuned to a lower frequency than the received signal's resting frequency. (The receiver could have been tuned to a higher frequency.) The resting frequency is at the center of the linear portion of the receiver's rf/i-f selectivity curve. When the

transmitter is modulated, its carrier will shift above and below the resting frequency at a rate determined by the modulating frequency. This is known as frequency deviation.

If the modulation is sinusoidal, one complete cycle will appear as in **B** of **fig. 1.** A zero-center millivoltmeter connected across the diode-detector load resistor of an a-m receiver tuned as in **fig. 1A** would indicate a pattern as shown in **fig. 1C** for the period of time depicted in **figs. 1B** and **C.** This corresponds to one complete cycle of the modulating frequency. Also, of course, it's the same as one complete cycle of the demodulated audio-frequency signal available at the receiver output.

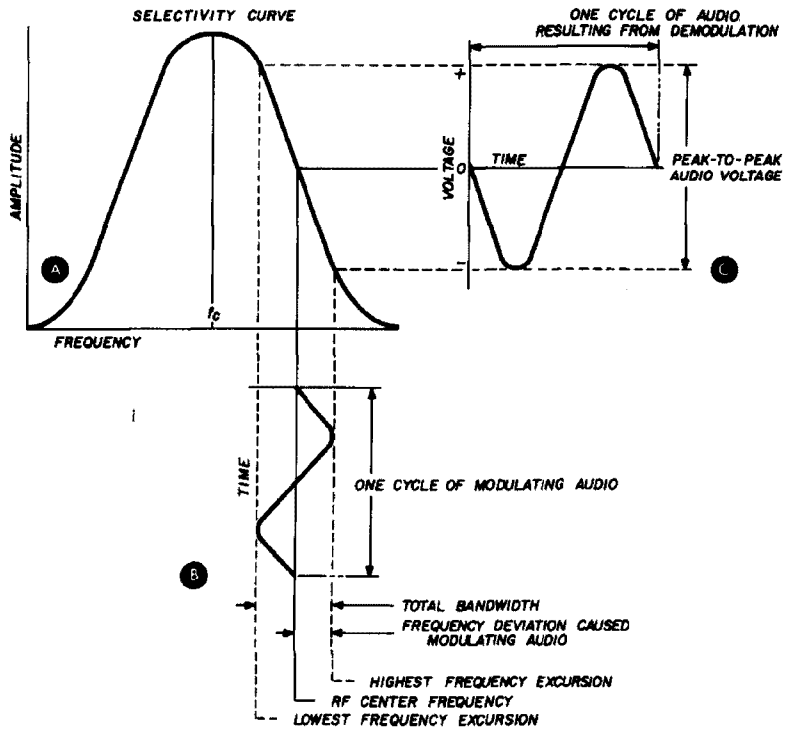


fig. 1. The relationship between transmitter frequency deviation and the demodulated audio signal. Usable audio signal is a linear function of transmitter frequency deviation.

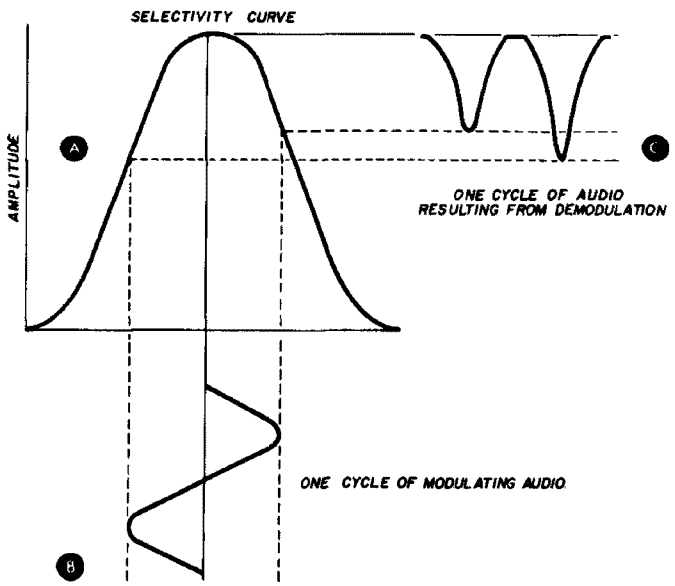


fig. 2. Effect of tuning the receiver to the center of the response curve. Note the flat top and extreme nonlinearity of the demodulated af signal.

If the input signal were redrawn to show only one-half the frequency deviation, the audio signal amplitude would be one-half its original value. From this you may draw the conclusion that an a-m receiver, correctly tuned and with a specific frequency response (selectivity) curve, will develop an undistorted audio signal that varies directly as the frequency deviation of the transmitted signal.

incorrect tuning

Two cases of incorrect tuning are depicted in **figs. 2** and **3.** The first shows a signal tuned too far down on the slope of the response curve. This has two effects. One reduces the amplitude of the demodulated signal; the other introduces some distortion in the audio waveform. If you tuned in the other direction; that is, too close to the maxi-

mum-signal (rf) portion of the selectivity curve, **fig. 2** shows what happens. Not only is the audio amplitude quite reduced, but a distressing degree of distortion appears.

So far, we've dealt only with a receiver having a rather broad, gently-sloping selectivity curve. Very few presently manufactured receivers for amateurs fall into this classification. Most such receivers have selectivity curves resembling that of **fig. 4,** with many coming quite close to the ideal flat-nosed, vertical-sided curve. If the curve has a reasonable slope, very **narrow** frequency deviations can be received quite well, but any wide deviation or incorrect tuning will cause serious distortion.

summing up

Slope-detection characteristics may be summarized as follows:

1. For other than narrowband fm, an a-m receiver selectivity curve should have a shallow slope. Its major portion should be free of abrupt change. Ideally, it should be straight.

2. If the receiver has a straight slope for a major portion of its selectivity curve, the receiver should be tuned so that the resting frequency of the incoming fm signal is in the center of the linear portion.

3. Thus tuned, the amplitude of the demodulated audio signal will vary directly as the deviation; the greater the deviation, the louder will be the audio.

4. Incorrect tuning (to either side of optimum) causes distortion.

5. Even with optimum tuning, excessive frequency deviation (for a given slope) will cause distortion.

6. For acceptable results, the transmitter frequency deviation must be compatible with the receiver slope characteristic. In other words, just any receiver will not be satisfactory with just any transmitter.

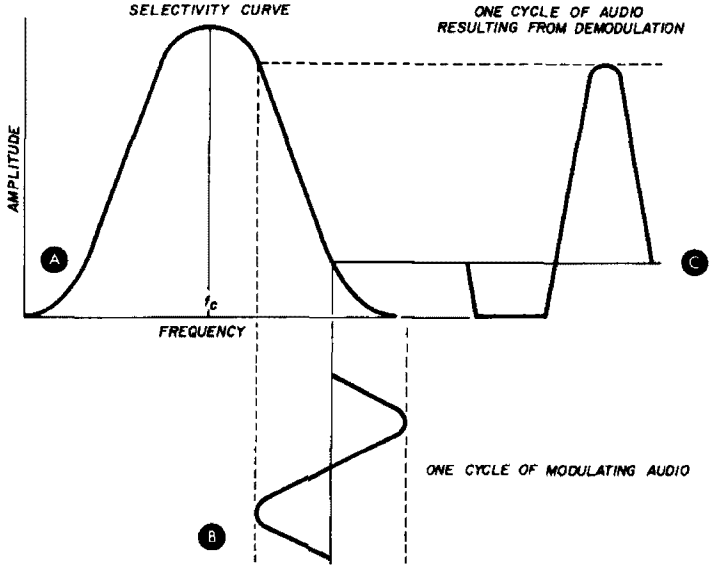
7. The more selective the a-m receiver (that is, the more nearly its selectivity curve approaches the flat-nosed, straight-sided shape) the less likely it will receive fm without distortion.

the discriminator

Let's now consider the discriminator, which is one of the most common means of converting fm to audio. **Fig. 5A** is a plot of voltage developed across a load resistor in the output of a discriminator. An rf signal is introduced on the y axis, **fig. 5B.** If its frequency is constant, no voltage will be produced. If, however, you vary the frequency of the incoming signal, you'd find an output voltage whose magnitude and polarity are related to the deviation. The demodulated signal amplitude is **not** related to the amplitude of the incoming signal. Note that the curve is linear over a major portion, which is the useful part. The nonlinear part would result in distortion.

One important fact must be thoroughly understood at this point. The curve of **fig. 5** is the over-all rf performance characteristic. It is not the discriminator output versus discriminator input, but the **antenna input** versus discriminator output. The **total** receiver selectivity curve must be considered if dis-

fig. 3. What happens when the receiver is tuned too far down on the response curve slope. Amplitude is reduced and distortion is introduced.



tortion is to be avoided. If the selectivity curve, up to the discriminator input, isn't flat-topped over a frequency range greater than that of the discriminator's linear portion, distortion will occur when the incoming signal has a bandwidth exceeding the flat-nosed part of the rf/i-f selectivity curve.

Most good a-m receivers must have their i-f strips broadened to receive fm for the reasons stated above. The exception to this is receiving narrowband fm. Most vhf fm doesn't fall into this category, however.

the transmitter

For compatible fm, receiver and transmitter must be matched. The exact task is to design an fm transmitter that conforms to predetermined standards. These usually apply to frequency deviation and its linearity.

Deviation often depends on some arbitrary decision. For example, the FCC makes a ruling, a group of amateurs concur on a standard for their net, or a MARS director issues a bulletin. Once the parameter has been set, it's up to the amateur to conform. Unless you have a commercially built deviation meter, conformance can present a real problem. You might obtain the services of a specialist in the commercial fm field. Get him to adjust your transmitter, then leave the controls alone. Especially, don't take the advice of well-meaning hams who may not know all the answers.

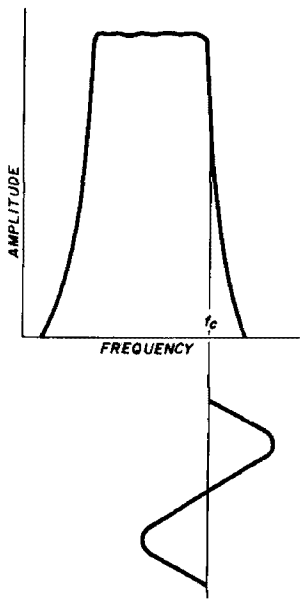


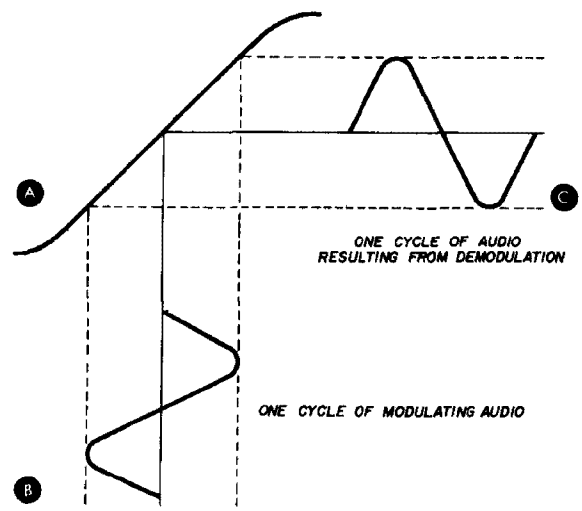
fig. 4. Typical response curve of many commercially manufactured amateur receivers. Tuning is generally critical, however, if the fm signal has a rather large frequency deviation.

dc method for deviation adjustment

To do this properly, you should have a peak-reading vtvm and a precision frequency meter. A steady voltage of one polarity is applied to the frequency-shifting circuit. The voltage is changed in small increments, the resulting frequency change is noted, and a plot is made of frequency deviation versus voltage. The polarity is reversed, and the procedure is repeated. The two plots should coincide exactly.

From these plots, the exact peak-to-peak input voltage is found that produces the

fig. 5. Typical fm discriminator curve. The demodulated af voltage amplitude is a function of the frequency deviation, not the amplitude of the input rf signal. Note the linearity, which gives undistorted output.

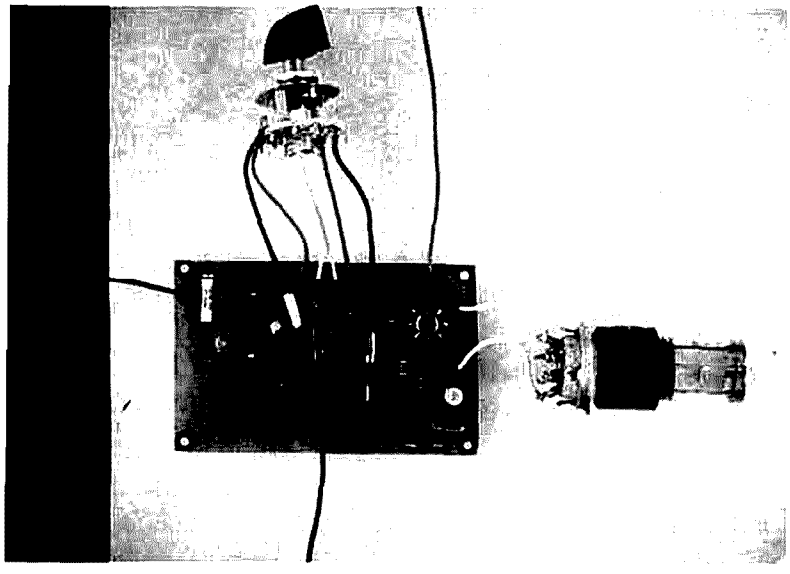


specified bandwidth. Then the audio circuit is adjusted for the same peak-to-peak voltage, and the control is locked. This procedure also takes care of the second standard: deviation linearity.

closing note

A review of the amateur literature on converting to fm has, with few exceptions, overlooked the considerations treated in this article. I hope the ideas presented here will encourage a more straightforward approach to the adjustment and operation of fm equipment.

ham radio



simple frequency-divider calibrator using mos ic's

Mosfet's have been used in many amateur rf amplifier designs—here's a different application

Hank Olson, W6GXN, P. O. Box 339, Menlo Park, Calif.

The metal oxide silicon field-effect transistor (MOSFET), also called MOST, is now available at moderate cost after extensive development. One of its more common applications is in the computer field, where it is being used in large-scale integration (LSI) circuits. The development of LSI arrays has brought forth several varieties of the MOST. For example, you can now obtain a single-chip device with a 1024-bit memory storage.[1] Or you can buy a MOST chip as simple as a three-input gate in one TO-5 package.[2]

Amateur ingenuity has produced MOST converters and other receiver front-end designs.[3-7] This article is another contribution to amateur application of solid-state devices —a frequency calibrator using several MOST's to produce a 100-kHz signal plus four submultiples. Construction is not dif-

ficult, but you should observe experience-proven techniques for handling and mounting the devices. Some tips on this are given in the following discussion.

the oscillator-buffer section

A three-input NOR gate (Philco pL4GO4) is used in this circuit. It could be termed a MOS array, since it consists of three p-channel enhancement-mode MOST's on one chip. Its internal circuit is shown in **fig. 1.**

The gate and drain leads of each transistor are brought out separately on the eight-lead TO-5 package. The sources are common and are tied to pin 6. The remaining pin (8) is connected to the substrate. Since these MOST devices are p-channel types, the power supply (E_{DD}) must be negative; −12 volts are used. Each MOST operates in the enhancement mode. This means that I_{DSS}, which is the drain current flowing when gate and source are connected, is nearly zero. Any gate must be negative with respect to its source before conduction begins. This is one of the nice features of enhancement-mode MOST's. They are self-biasing, like the zero-bias triodes that are so popular in amateur linear amplifiers.

Two of the MOST's in the pL4GO4 are wired as an astable multivibrator; a crystal and capacitor in series replace the usual cross-coupling capacitor. This crystal-controlled multivibrator then becomes our frequency standard. The remaining MOST is used as an isolation amplifier.

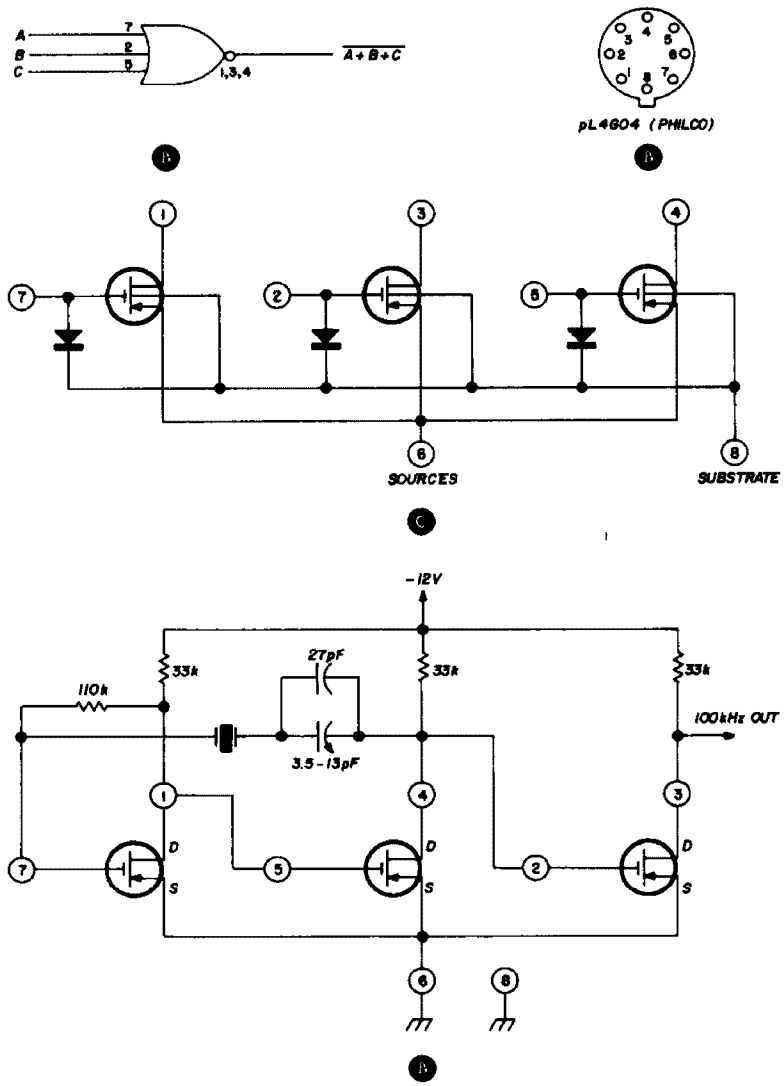


fig. 1. Three-input NOR gate used as the oscillator-buffer. Logic and pin diagrams are shown in (A) and (B). Internal circuit (C) shows protective snubbing diodes. The first two MOST's in (D) form the 100-kHz standard.

divider and shaping circuits

Following the oscillator-buffer is another MOST IC containing four binary divider stages and four inverting buffers (one for each divider), **fig. 2.** This IC is specifically designed for electronic organs, so its commercial price is low. A data sheet is available from Motorola.[8]

The desired output (100, 50, 25, 12.5 or 6.25 kHz) is applied to a Schmitt trigger. This circuit shapes the signal to a near-per-

fect square wave, thus increasing harmonic content.

Whereas the p-channel MOST's (pL4GO4 and MC1124P) are relatively slow, the n-channel MOST and npn bipolar transistor in the Schmitt trigger will toggle above 100 MHz.

Solid-state Schmitt triggers ordinarily use two bipolar transistors; this one uses one MOST and one bipolar device. The MOST in the input of the Schmitt trigger allows it to be driven by the pL4GO4 or MC1124P, which have high output impedances. The bipolar transistor on the output side has a much better saturation characteristic than a MOST, providing an output pulse approaching an ideal square wave. Thus by combining transistor types, one can achieve better performance characteristics than by using only one pair of either type.

construction

Those who haven't used MOST's seem to be either completely unaware of their fragility or are afraid to touch one for fear of wrecking it. This is understandable, because early MOST's were indeed susceptible to gate burnout from electrostatic charges. This can occur from your body or from the polyfoam package material. However, most MOST IC's today have snubbing diodes between gate and source to prevent voltage buildup between these elements. The IC's in this article are of that type. The 3N128 in the Schmitt trigger is **not**, however, and it was handled by wrapping a fine wire around its leads. The wire was put on before the shorting collet was removed. The 3N128 was soldered into the circuit last, and the 110-kilohm gate-to-source resistor was soldered into the circuit before removing the short-

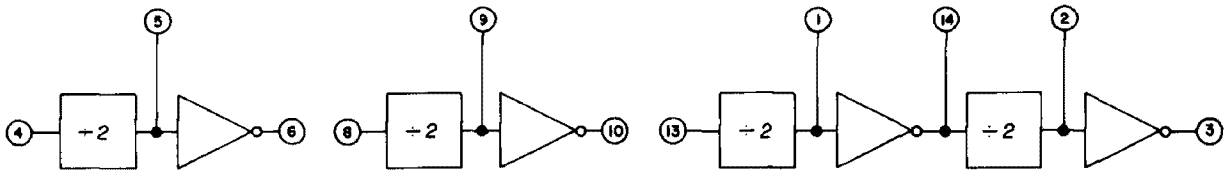
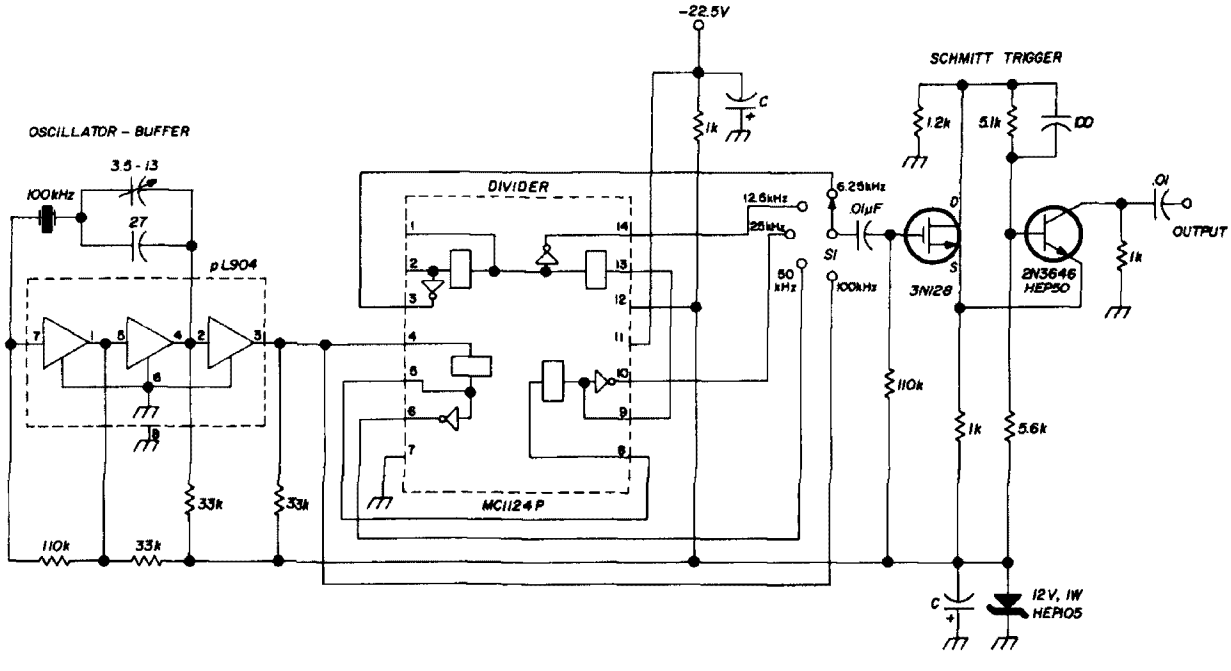


fig. 2. The MOST IC used as a frequency divider. Logic has been modified for this purpose to obtain the fundamental 100 kHz and four submultiples.

fig. 3. Complete calibrator circuit. The 3N128 should be handled with care to avoid static charges from building up between source and gate; see text.



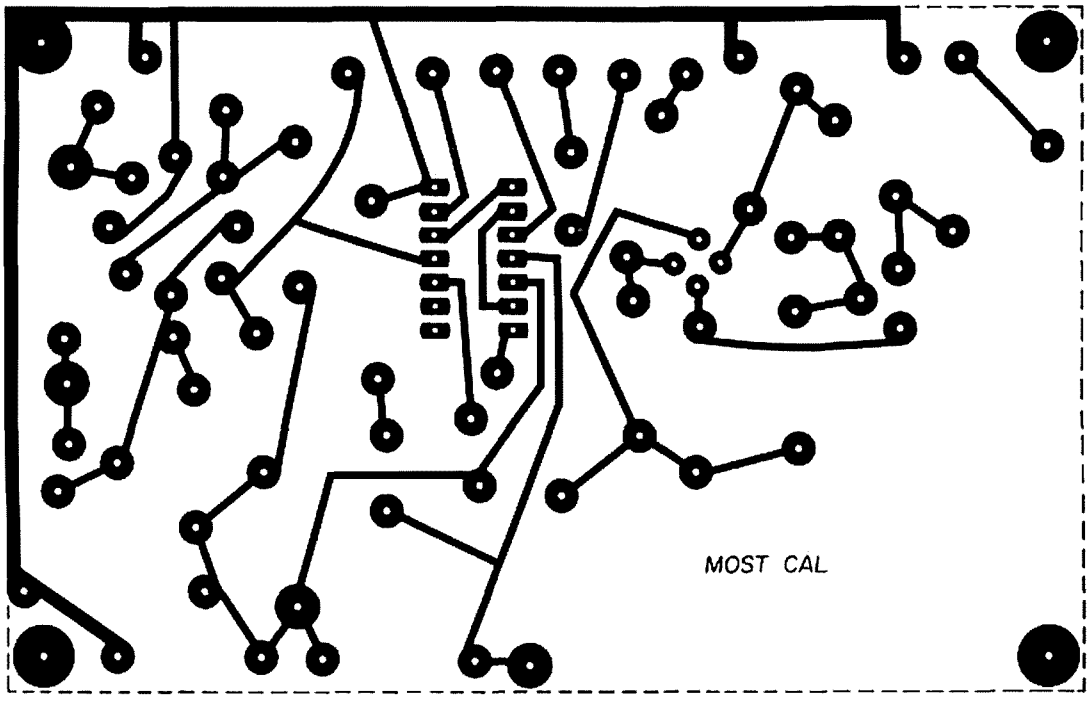


fig. 4. Etched circuit board for the frequency-divider calibrator.

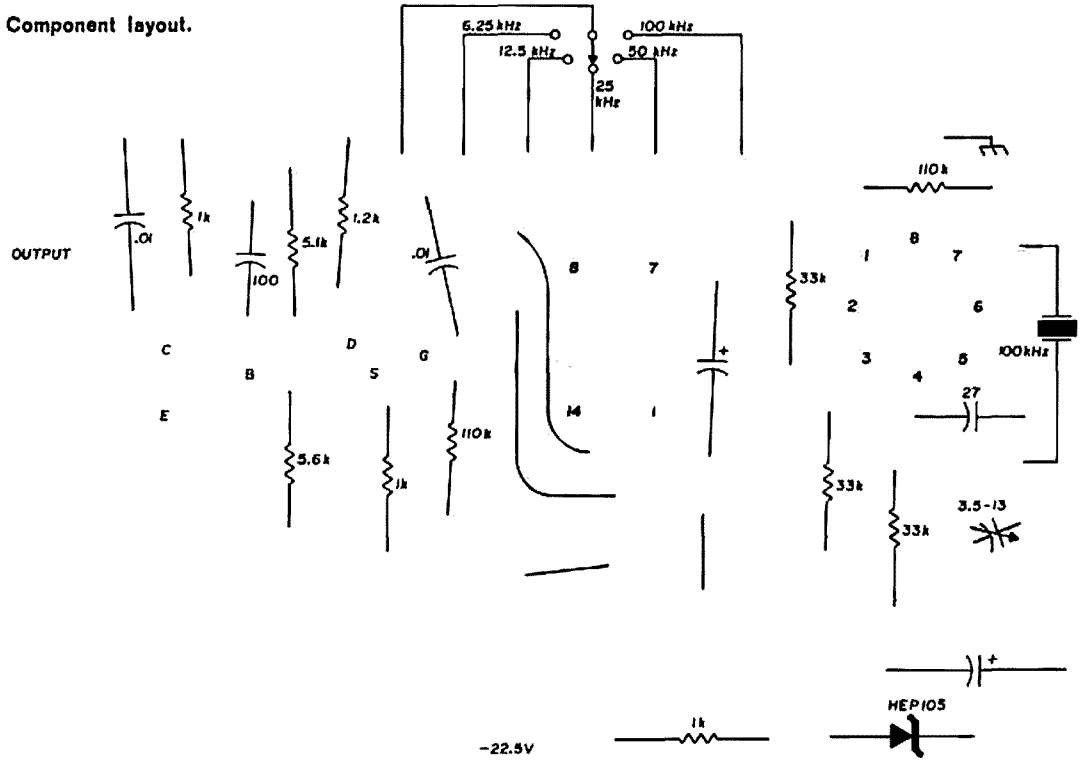
ing wire.

The complete calibrator schematic is shown in **fig. 3.** The pL4G04 and MC1124P were mounted in sockets, but sockets aren't absolutely necessary. If you'd like to include this refinement, the sockets I used were Nugent LP5178 and Methode M1141. The crystal is a Bliley glass-mounted type.*

The frequency-adjusting capacitor is made by Trush, Inc., part number 7S-TRIKO 02.

* Bliley BG9D, available from Red Johnson Electronics, 440 Pepper Street, Palo Alto, California. Price $5.00 pp.

Component layout.



The circuit board, **fig. 4,** is designed especially to fit this particular capacitor. A ready-made board, complete with parts, can be obtained if you don't wish to make your own board.**

A single-pole, 5-position wafer switch, S1, selects the desired frequency. A suggested source is Mallory, part number 3215J. The two bypass capacitors, identified as "C" in **fig. 3,** aren't critical. Tantalytic or Mylar low-voltage capacitors (0.1 to 20 μF) will work fine.

A circuit for a regulated power supply is shown in **fig. 5.** The calibrator will operate for many hours on a battery supply, however. A Burgess number 4156 may be used.

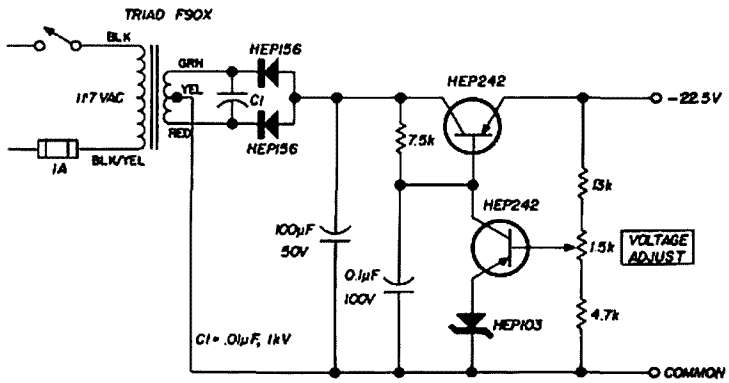
predictions for the future

The first cut at the calibrator design used a Philco pL4S10 ten-input NOR gate, which has 22 leads on its small package. Only three of the ten gates were used. The complexity of such off-the-shelf IC's is an indication of how fast LSI components are becoming available to amateurs. The cost per gate of this device is lower than that of the pL4GO4, even though the flat-pack styles are generally more expensive than the TO-5 package.

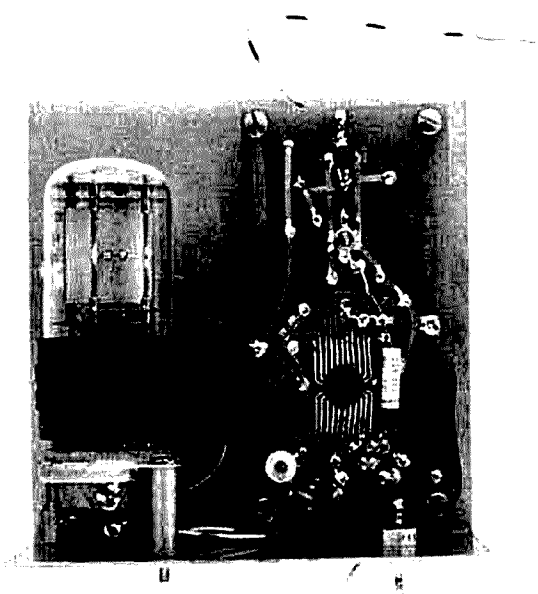
Prices of MOST integrated circuits should decrease as competition increases among manufacturers. The electronic organ industry indicates that something on the order of 40¢ per divide-by-two stage is the economic break point to justify the use of microcircuits. It's almost a certainty that this figure will be realized next year. This should make the price about 80¢ apiece for amateurs for 1 to 99 pieces. This should undercut the price of even the inexpensive RTL IC's, which have gained popularity among amateurs.[9]

** Project Supply Company, P. O. Box 555, Tempe, Arizona 85281.

fig. 5. Suggested power supply for the frequency calibrator.



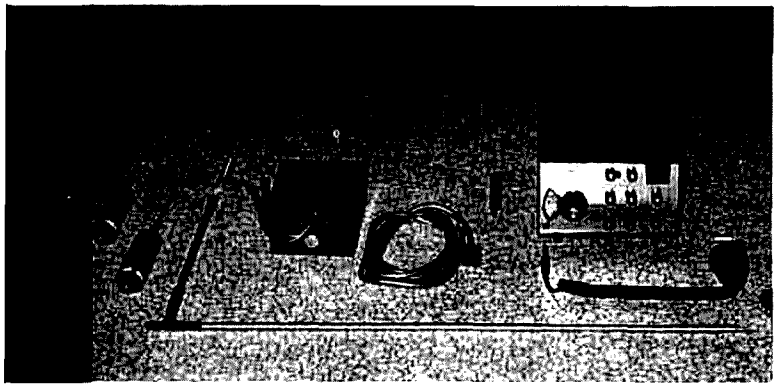
Original version of the calibrator, using a Philco ten-input NOR gate. Only three gates were used for this application.



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2. Data sheet for Philco pL4G04, M-178, May, 1967, Philco Microelectronics, Blue Bell, Pennsylvania.
3. Mike Goldstein, VE3GFN, "Bandswitching FET Converter," *ham radio,* July, 1968, p. 6.
4. Don Nelson, "The WB2EGZ Six-Meter MOSFET Converter," *ham radio,* June, 1968, p. 22.
5. Don Nelson, WB2EGZ, "MOSFET Converter for Two Meters," *ham radio,* August, 1968, p. 22.
6. Don Nelson, WB2EGZ, "220-MHz MOSFET Converter," *ham radio,* January, 1969, p. 28.
7. Hank Cross, W1OOP, "Converting a Vacuum-Tube Receiver to Solid State," *ham radio,* February, 1969, p. 26.
8. Data sheet for Motorola MC1124P, PS-34, September, 1967, Motorola Semiconductors, Box 955, Phoenix, Arizona.
9. H. Olson, "Micrologic for Nonlogical Users," *73,* June, 1967, p. 58.

ham radio



putting together a mobile installation

A recipe for summer fun

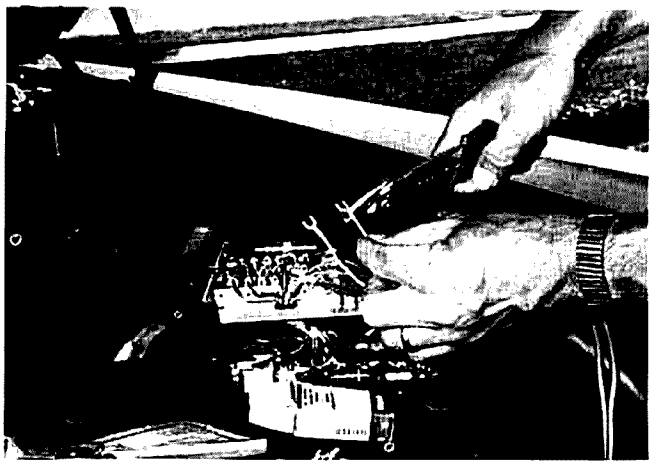
Bob Ruyle, WØFCH, 420 Steinway Road, Lincoln, Nebraska, 68505

ingredients:

1 Galaxy transceiver
1 mobile power supply
1 mobile mounting bracket and speaker
1 Antenna Specialist M3D body mount
1 Hustler foldover mast
1 Hustler resonator spring
Waters loading coils and kilowatt tips

Route the power-supply cable from the mounting position of the transceiver to the area near the battery where you're going to mount the power supply. Keep the power supply away from areas that retain heat; I mounted mine up toward the front and even had to reposition the horn. Route the cable neatly through cable hangers.

Take the bottom plate off the power supply and connect the power cable as shown in the instruction manual. Check all connections before sealing the unit up again. Install the power supply in the spot you've chosen. Route the dc primary cables to the battery using the **shortest** path possible. Primary cables should be no longer than 4 feet. I ran mine in the cable channel between the radiator and air conditioner.



Finishing the wire ends.

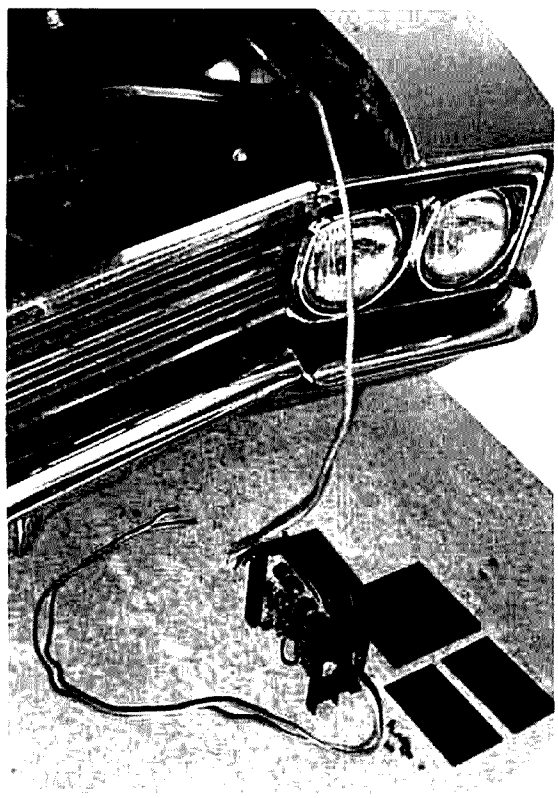
Take a 20-foot length of good quality RG-58A/U coax (or RG-8A/U if you have the space) and route it from the front seat area to the back of the car where the antenna is to be mounted. You can remove paneling and covers to hide the coaxial cable—in the photo the cable is routed underneath the carpeting and behind the side panels. Be careful not to cut or pinch the cable when replacing the panels and protectice covers.

Solder the uhf connector to the antenna lead; check the cable with an ohmmeter to make sure you haven't shorted it out.

Now lay out the pieces of the M3D antenna mount and attach the center conductors and ground lead. Remove all the hardware and mark the fender for drilling. A Greenlee punch is handy for the large round center cutout.

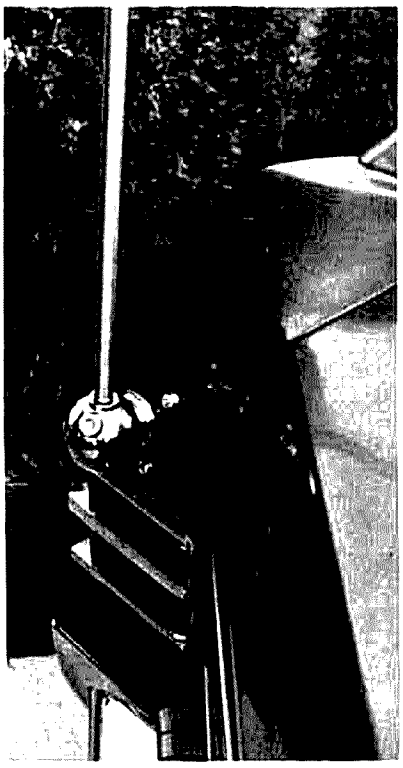
Punching the antenna-mounting hole.



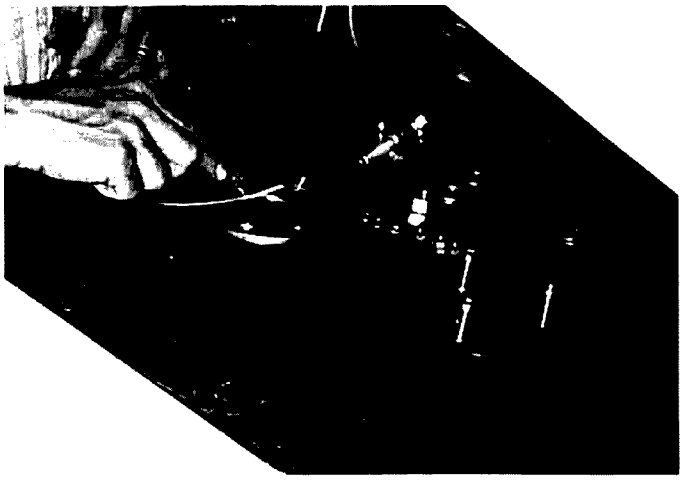


Above, mobile power supply and cable. Photo below shows routing of coaxial cable under side panels.











After the mounting hole is punched in the fender, put the mount together and screw in the mast, resonator spring, loading coil and tip.





Install the mobile mount.

Put the M3D antenna mount together, screw in the mast, resonator spring, loading coil and tip. Your mobile antenna is now ready for tuning.

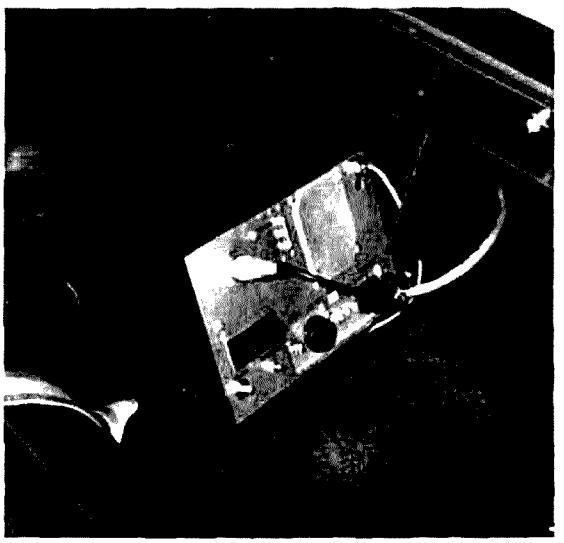
Next, install the mobile mounting bracket. Be sure to try it out with the transceiver attached before marking any mounting holes—try all dials, switches, ash trays and the glove compartment door to make sure they don't interfere with the chosen transceiver location.

Connect the speaker, dc power cable and the antenna lead-in. Slip the transceiver into the mobile mount and run a ground line from the transceiver to the car body.

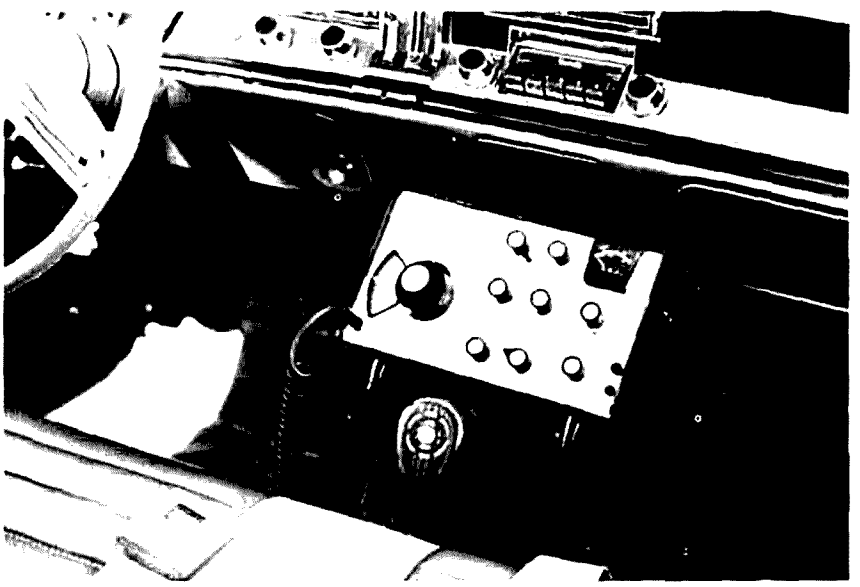
I mounted the speaker enclosure and the microphone to the bottom of the transceiver cabinet. This puts everything within easy reach when driving.

After tuning up the antenna, I fired up the installation and checked power output, swr and current drain under full load. Then I gave a call on 40 meters, and got a solid report from WØCUS near Colorado Springs. The recipe was complete. I would like to thank Galaxy electronics for their help in putting together the ingredients for this recipe.

ham radio



Hook up all the cables.



The completed installation.

a new multiband quad antenna

This driven array features several improvements over conventional quads for three-band operation

Werner Boldt, PhD, DJ4VM, Am Zuckerberg 4773, Körbecke, West Germany

The cubical quad antenna has a reputation among amateurs for being an excellent DX antenna. Even after accounting for claims of over-enthusiastic quad users, the fact remains that the cubical quad is an excellent low-angle radiator, especially where height is limited.

Several ways to build multiband quads have been proposed. Some have proved to be practical and have become popular. All, however, are compromises in one way or another. Before describing the DJ4VM quad, which has been designed to exploit the best features of the quad antenna, I'll review the most popular designs and discuss their disadvantages.

the multiband quad

The most popular multiband quad design consists of nested loop elements on a common supporting structure. Each antenna for each band is supposedly independent, and all are fed with a common transmission line using a gamma matching system or other tricks. The disadvantage here is that the size of the structure, which is determined by the lowest-frequency antenna, is not fully realized on the higher-frequency bands. In the common 20/15/10-meter antenna, for example, the 10-meter loop uses only about 25 percent of the available area.

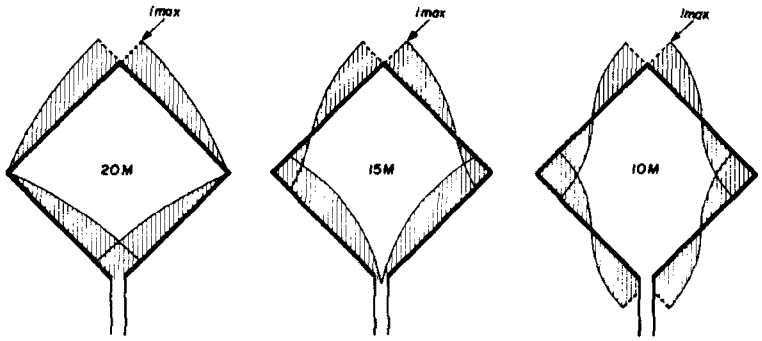


fig. 1. Current distribution in conventional quad elements (cut for 20 meters). Note displacement of current loop with frequency.

forward field

The forward gain of each of the three antennas is expected to be the same as that of separate antennas of the same loop size. In concentric arrangements, however, the antennas interact. They exhibit spurious responses off the sides and to the rear. It is almost impossible to obtain more than 15 dB front-to-back ratio in the 15-meter antenna when loops are mounted concentrically with 10- and 20-meter loops.[1]

Investigations of single-loop arrangements have been made to simplify the antenna structure and take advantage of the given structure. The loop is usually fed in the lower-half portion; tuning and matching is achieved by stubs or lumped constants.[2,3]

If we consider the current distributions on different bands **(fig. 1)**, we can draw the following conclusions:

1. On 20 meters, currents in the upper and lower halves of the loop are **in phase.** The loop thus acts as a broadside array, ensuring low-angle radiation in the desired direction.

2. On 10 meters, the currents in the upper and lower halves are **180 degrees out of phase.** The antenna performs as an end-fire array. Enhanced radiation is parallel to the plane of the loop (vertically in this case). This doesn't do much good in long-haul DX work.

3. In-phase current distribution prevails on the lowest-frequency band; rapid deterioration of performance occurs with increase in frequency.

By introducing stubs, lumped constants, and switching schemes, the proper phase relationships can be obtained on several bands. On ten meters, for instance, the loop can be opened at the top corner, thus converting the antenna into a common bi-square, or so-called inverted rhombic.[4] These designs are feasible but not too practical.

the center-fed loop

To overcome the disadvantages described above, I have developed a center-fed loop element as shown in **fig. 2.** Symmetrical current distribution with respect to the horizontal axis is achieved on all three bands without switching or tuning the elements. Without introducing compromises, the leg length of one edge of the diamond may be between one-quarter and five-eighths wavelength. This will yield a usable frequency range of about 2.5:1 for a fixed loop. Exceeding the maximum length will result in spurious side lobes.

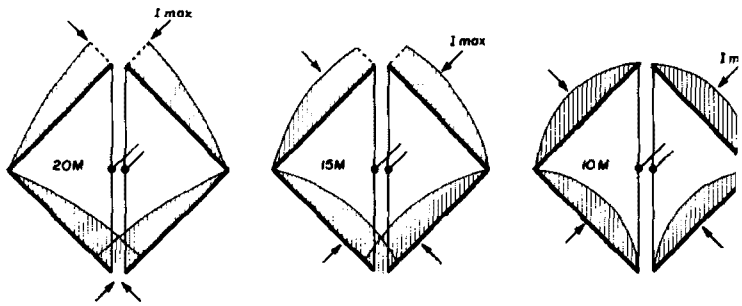


fig. 2. Current distribution in center-driven quad elements. Current loops are always symmetrical.

leg length

The leg length can be less than one-quarter wavelength, but performance will be degraded, part of which is because the current loop is no longer located in the center of each leg. It should be possible to

correct this with loading capacitors, as shown in **fig. 3.** However, this means accepting losses introduced by the capacitors.

In general, leg length should be as long as possible, observing the 5/8-wavelength limit. Optimum leg length for a triband antenna would be 5/8 wavelength on 10 meters, or about 22 feet. However, a leg length of 17 feet (1/4-wavelength on 20 meter) would still be a good design, exhibiting only slightly less overall gain.

By settling for less than optimum gain, and sacrificing performance on 20 meters in particular, it is possible to get an element of this kind on a structure that would ordinarily support only a 15-meter quad element and obtain an antenna usable from 20 through 6 meters (12-foot leg length) with excellent performance on 6 meters.

with 17 to 22 feet on a leg. The switching can be done either manually or remotely. If you use variable capacitors with a maximum capacity of 50 pF, the coils should have the following approximate dimensions:

20 meters: 10 turns, 1 1/2-inch diameter
15 meters: 8 turns, 1 1/2-inch diameter
10 meters: 8 turns, 1 1/4-inch diameter

fig. 4. Basic two-element arrangement of the DJ4VM multiband quad.

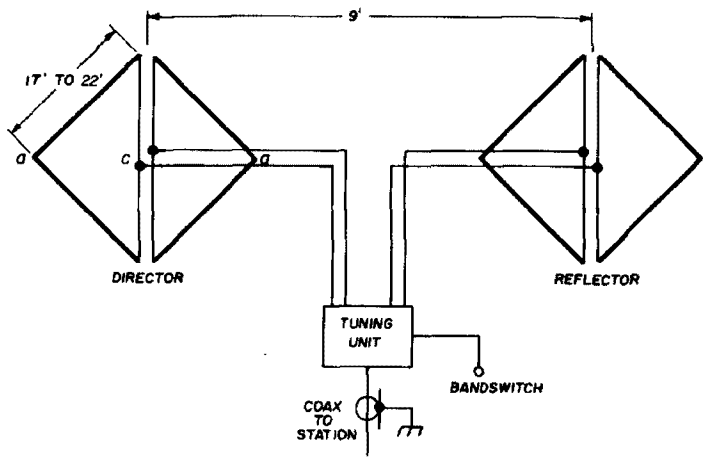
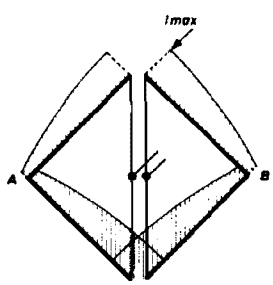


fig. 3. A miniquad element with loading capacitor at A and B.



the DJ4VM quad

The basic two-element arrangement is shown in **fig. 4.** Two designs were investigated: parasitic reflector and driven reflector. As shown later, the latter has definite advantages. The main one is the greatly reduced need for making a careful compromise in choosing the spacing between the two loops. This is achieved by carefully considering forward gain, front-to-back ratio and bandwidth on all three bands.

the parasitic version

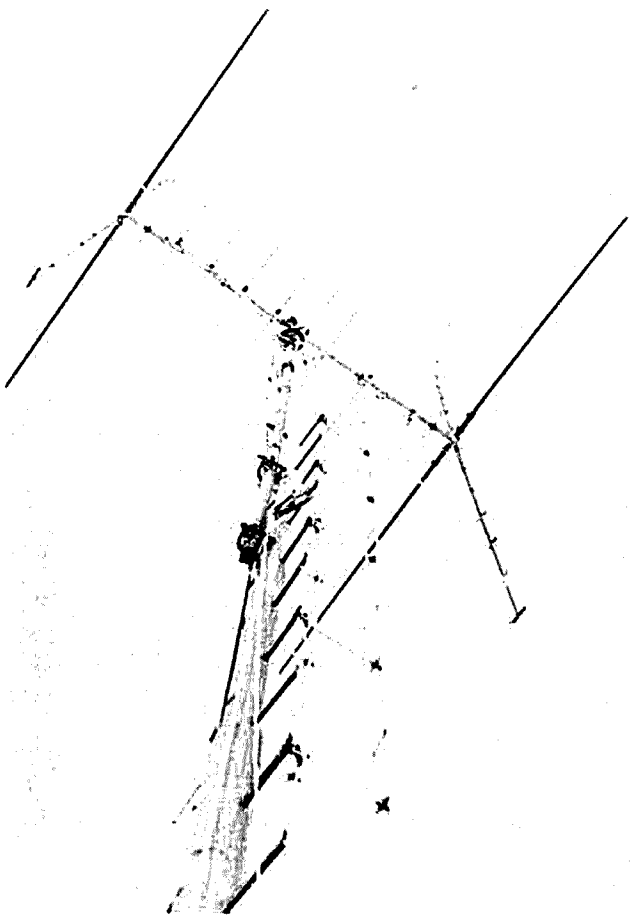
The tuning apparatus for this system is shown in **fig. 5.** Element spacing is 9 feet,

All are spacewound, of course, and preferably self-supporting. Instead of tapping the coils, the coaxial cable could be coupled with a link. Capacitor Ck should be used to tune out leakage inductance.

C1B should be adjusted for maximum front-to-back ratio. Thereafter the setting of C1A should be rechecked and corrected if necessary.

The tuning unit for the dual-driven antenna is shown in **fig. 6,** which is very similar to **fig. 5.** The two elements are driven out of phase by about 180 degrees. The lead length from the branch of the incoming coax to the switch contact, and from there to the coil taps, should be kept reasonably short to eliminate need for transformation. The tap point must be determined by using an swr meter. The approximate tap locations will be as follows (turns counted from the center tap):

20 meters: 2 turns
15 meters: 1 1/2 turns
10 meters: 1 turn



Multiband quad installation at DJ4VM.

For a better adjustment of front-to-back ratio, capacitor Ck can be used. If you do this, the feed point on the reflector coil will change to the opposite side of the center tap (Lb'). It will also be slightly farther away from the center.

Final tuning should be done with a field-strength meter and swr bridge while tuning C1B in, C1A out, and Ck for maximum attenuation to the rear. It's important to take the time and trouble to do this properly if you want to realize the best performance from this antenna.

Taps for the feed point should be chosen so that C1A and C1B resonate at about the same capacitance (with the transmission line connected and disconnected).

field test results

I don't have any quantitative measurement figures on forward gain. Instead of unsubstantiated claims resulting from wishful thinking, however, I'll give some field test results.

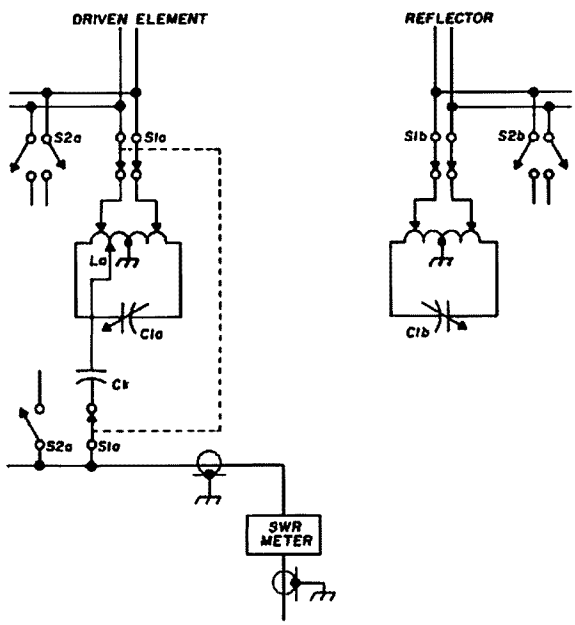
According to reference 5, the beamwidth of an antenna is an indication of forward gain. The half-power points of the antenna described were, from field tests:

20 meters: 50 degrees
15 meters: 40 degrees
10 meters: 30 degrees

Bear in mind this data was taken from my location and will not necessarily be the same for yours. Nevertheless, even under less-than-optimum local geometry, these beamwidths certainly are competitive with most 3-element Yagi antennas. The increased gain (narrower beamwidth) on the higher frequencies results from the larger radiator size, which is a direct consequence of the design principle.

The measured response on 15 meters is depicted in fig. 7. The front-to-back ratio was more than 40 dB. Without readjustments in the tuning unit, however, this value can be maintained over a relatively narrow frequency range. Fig. 8 shows swr, front-to-back ratio and relative gain for the 15-meter band.

fig. 5. Tuning circuit for the parasitic antenna. Switching can be either manual or remotely controlled. Capacitor Ck tunes out leakage inductance.



advantages of the driven antenna

The main advantages of this antenna, compared with other beam antennas in general, are:

1. Utmost flexibility when scrambling for DX contest points.
2. No need to align antenna elements; all tuning is done remotely with the circuits in the tuning unit.
3. Lower angle of radiation for a given height.
4. Instant 180-degree switching of the beam degrees by switching off the reflector relays. (How long does it take to swing a Yagi from East to West?)
5. Less lateral space required than for a full-sized Yagi, colinear, long-wire and most other horizontally-oriented beams.

a closing note

There's nothing secret about the diamond shape. It was chosen because the vertical feedlines could be easily attached to the fiberglass poles. Other shapes such as squares, circles (for vhf), or the Swiss quad[6] could be adapted as centerfed, multiband quad elements.

The all-driven quad has been tested for some time and has given excellent results. Neither the antenna height (40 feet) nor the location (in a valley) can explain the performance it has given.

fig. 7. Measured horizontal radiation pattern of the all-driven array. Design center is for the 15-meter band.

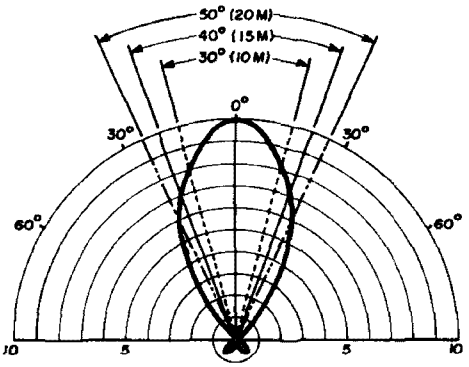


fig. 8. Standing-wave ratio, front-to-back ratio, and relative gain versus frequency. The data was taken on the 15-meter band.

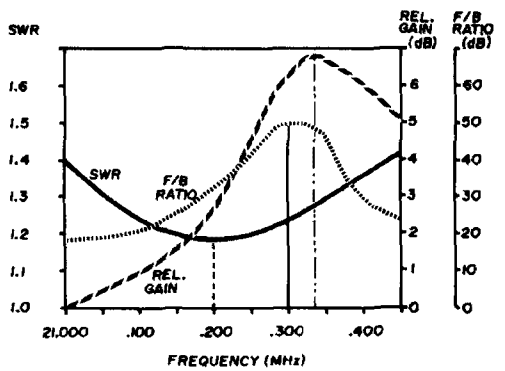
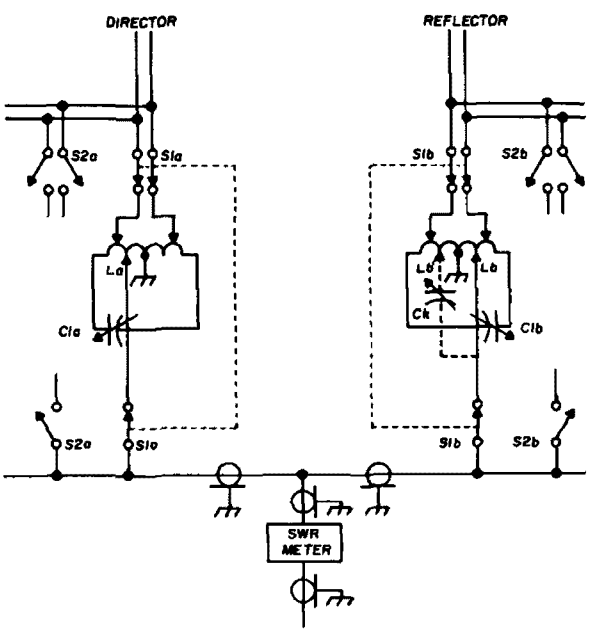


fig. 6. Tuning circuit for the all-driven array. Points La, Lb and Lb' are connected to the coax center conductor. The dashed line between Ck and the coil is an alternative.



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3. H. F. Rueckert, VK2AOU, "Dreiband-Eindraht-Cubical-Quad-Element," *DL-QTC* (Germany), 1968/4.
4. Byron Goodman, "Inverted Rhombics and Biconical Beams," *QST*, June, 1949, p. 42.
5. W. I. Orr, W6SAI, "Beam Antenna Handbook," Radio Publications, Inc., Wilton, Conn., 1955, p. 31.
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a new cw monitor

The versatile IC appears again— this time in an rf-actuated keying monitor featuring the low-cost μL914

John J. Schultz, W2EEY, 40 Rossie Street, Mystic, Connecticut 06355

Rf-actuated keying monitors certainly aren't new. They've been described in many articles, and there is even a commercial version available. However, the use of simple, inexpensive IC's, plus a different method of audio keying, make the monitor described here a worthwhile and interesting project.

The main advantage of rf-actuated monitors, of course, is that no direct connection to the transmitter is needed: the monitor uses internal battery power and is actuated by the rf field from the transmitter. Another advantage to such a monitor, which is less precisely explainable but nonetheless very real—as most users of such monitors will testify—is the clean keying that results. Clicks and other distortion present with most directly actuated monitors are completely eliminated.

features

This rf-actuated monitor is unique in several respects. It uses only IC's and is a very simple project for anyone who would like to get acquainted with these devices. The IC is the Fairchild μL914. It is becoming one of the most versatile IC's any amateur can use. As its designation (μL for micrologic) implies, it was developed for digital logic circuits but it serves very well as a linear IC that will amplify continuously varying af or rf signals up to several megahertz. It's available from large mail-order supply houses for less than a dollar.

Another interesting feature of the monitor is that the transmitter rf field is not used to activate a transistor switch to turn on an audio oscillator, as is commonly done. The audio oscillator in this monitor runs continuously. The rf field controls a μL914 that is used as an enabling gate or switch. Audio passes through the gate only when the rf field closes or "enables" the circuit. The distortion and transients which occur when an audio oscillator is initially turned on with each transmitted keying character are thus avoided.

Particularly clean-sounding audio keying results from using this method. The audio oscillator shown in the circuit (another μL914) is of fixed frequency; but if you have another audio oscillator available, perhaps of variable frequency, it could be used in place of the oscillator shown simply by feeding it directly into the enabling gate. Exactly how this can be done is covered later.

circuit description

The μL914 consists of a pair of dual grounded-emitter transistor stages. External connections determine how the stages are used in any particular circuit function. Physically, the μL914 appears as an ordinary epoxy-cased transistor, except it has eight leads. The numbering of these leads follows the scheme shown in **fig. 1.**

It's particularly important to note that the **index** flat on the μL914 case denotes the

8th terminal lead. Anyone who is used to working with tubes will probably regard the index mark as denoting the 1st terminal lead, as I frequently did when first working with the μL914. The result, of course, can very easily be ruined IC's.

The monitor circuit is shown in **fig. 1.** The first μL914 is connected as a free-running multivibrator. The oscillation frequency with the circuit components shown is approximately 1 kHz but the frequency may be changed over a wide range by varying either the value of both resistors or both capacitors.

For variations over a small range—if you can't get used to the idea of hearing the same tone all the time—you can replace one of the resistors with a 20k-ohm potentiometer.

The second μL914 is actually connected to perform a logic function. In this case, however, the result is as though an spst switch, connected between terminals 1 and 7, were controlled by a dc-control voltage on terminal 3 (generated by rectification of the transmitter rf field).

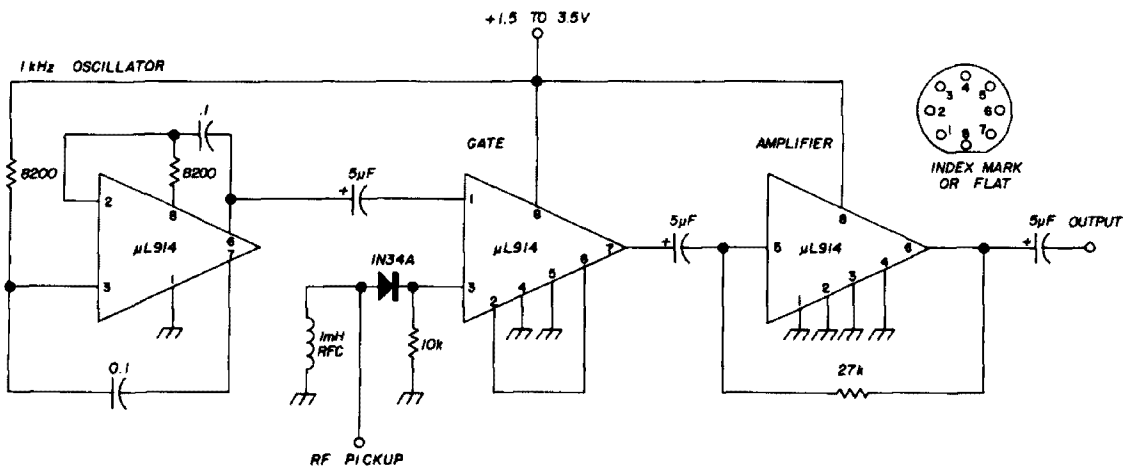


fig. 1. Schematic diagram of the integrated-circuit cw monitor.

operation

The 1N34A rectifies the rf pickup to produce a positive control voltage at terminal 3 of the second μL914. The rf pickup may be obtained by connecting a short piece of wire (about 1 foot) to the rf pickup terminal, which can be placed anywhere in the vicinity of the transmitter power amplifier.

The last μL914, which is used as an audio amplifier stage, actually makes very poor use of the capability of the μL914. Only one transistor in the μL914 is active as an amplifier. A single audio transistor can be used for this unit if desired. Surprisingly enough, the cost will be about the same as that using the IC.

The amplifier stage output is sufficient to drive a medium-impedance (400- to 600-ohm) headset with adequate volume. If loudspeaker operation is preferred, the last μL914 can be followed by a transistor power amplifier of conventional design.

No special precautions need be followed in construction, except perhaps to note that the leads concerned with the rf pickup circuit should be kept as short as possible. The circuit of **fig. 1** can be assembled on a 1-by-1½-inch piece of Vector board. The circuit can be mounted either inside a separate enclosure or inside the cabinet of your transmitter or transceiver.

The supply voltage is between 1.5 and 3.5 volts. The monitor will work well within this range, but at least 3 volts are necessary if the last μL914 drives a pair of headphones directly. The current drain is very low and can be handled for an extended period by two penlight cells in series.

ham radio

a combined digital and burst encoder

Selective call and tone-burst signalling provide enhanced fm operation

David M. Stahley, K8AUH, 6401 Brookside Drive, Cleveland, Ohio 44144

Sociologists tell us the world is suffering from overpopulation. Anyone who tunes across the amateur bands will certainly agree, and the "fm frequencies" of 52.525 and 146.94 MHz are no exception. The availability of commercial fm equipment at moderate cost has no doubt contributed to this condition.

Efforts to minimize operating problems and retain the enjoyment of fm operation have produced the techniques of tone burst and selective call. A combination selective-call—tone-burst encoder is described in this article for fm enthusiasts interested in improving station versatility.

Selective call is used to alert another station to your traffic, allows remote control, and provides specific address for autostart teletype. Tone-burst eliminates the need for constant monitoring of a crowded channel (by your wife, for example) to ensure receipt of traffic. It also allows more positive control of relay and repeater facilities.

features

The encoder described here, which can be used for either mobile or base operation, provides two or more selectable control tones, burst or digital command, and acts as a microphone preamplifier and multi-rig station control.

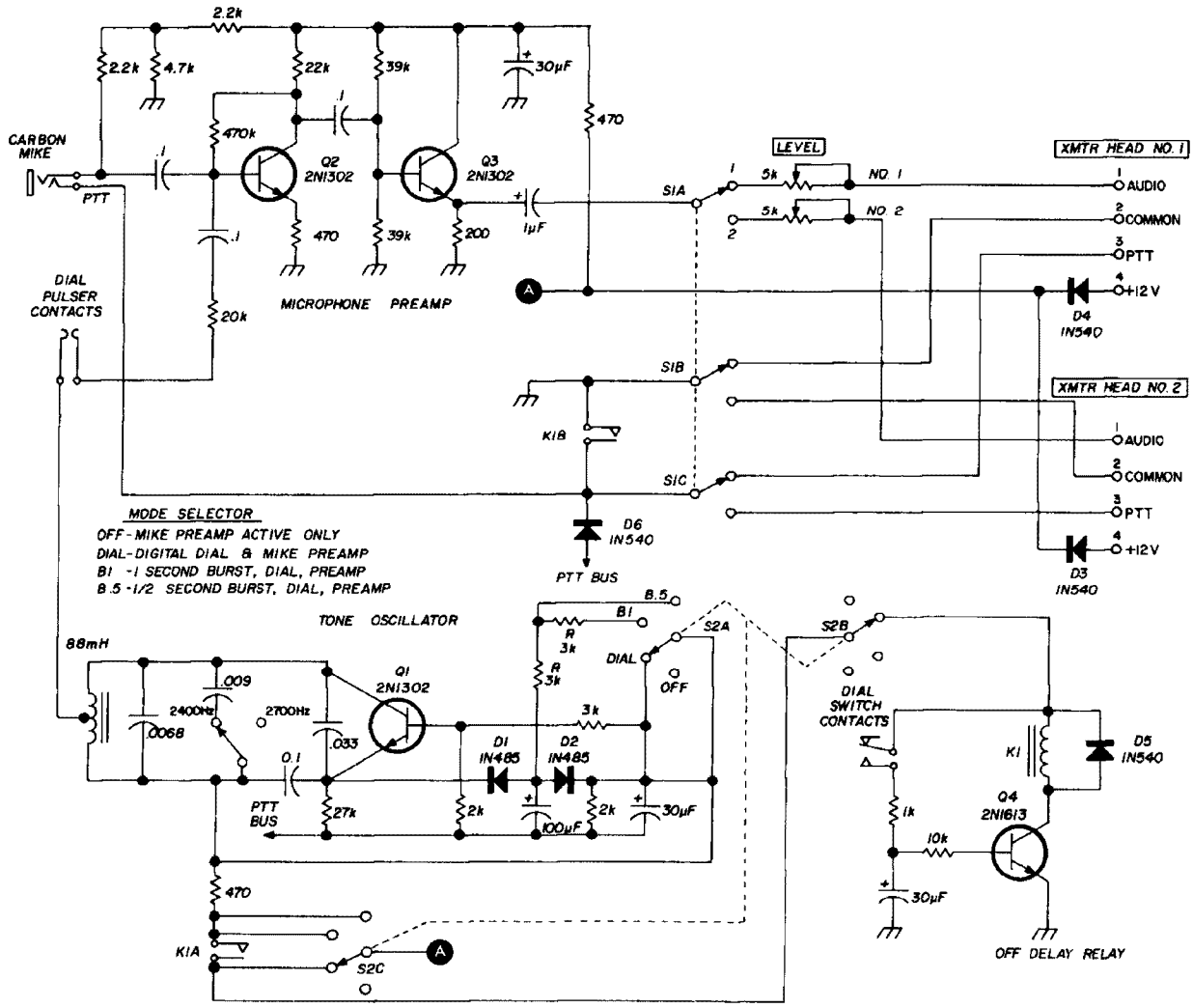


fig. 1. A schematic of a combined digital and burst encoder. K1 is a 12 Vdc relay.

A four-position selector switch provides:

1. A one-half-second tone burst at the beginning of each transmission.

2. A one-second tone burst as above.

3. Digital dial encoding and carrier-off delay.

4. Microphone preamp only.

The carrier-off delay feature (**fig. 1**) can be eliminated if you wish. You can then retain the dial-operated keying relay, and the selector can be left in the dial position without interfering with the push-to-talk (PTT) circuit.

digital dialing

During dial operation, the normally open contacts on the dial switch operate a relay that keys the transmitter and energizes a tone oscillator. If "off delay" is included, the carrier will remain on after the dial returns home. This allows operation of decoder alarms or other accessories.

The off-delay feature enhances multi-digit dialing. The dial-delay-off circuit can be any combination that will allow K1 to pull in immediately and drop out about 1 to 4 seconds after the dial returns home. This can be provided by a time-delay relay, a copper-slugged relay, capacitor and diode around the relay coil, or an RC-transistor timing circuit similar to that shown. K1 and Q4 comprise the delay-off circuit.

The digital circuit includes an oscillator

whose output is interrupted by a standard telephone dial. K1 provides the means by which the digital mode is enabled and allows automatic operation by simply dialing. No manual switching of audio lines or carrier switching is required.

the burst mode

Burst operation must be selected and will provide a tone burst at the beginning of each transmission. This mode lends itself to burst entry of a base or repeater receiver, and provides relief from the need for constant monitoring. It also allows a more positive control of such facilities. The burst is accomplished by squelching off the oscillator with an RC network and switching diode D1.

Burst duration is determined by the value of R in the charging circuit. D2 provides a discharge path for C when the microphone is released at the end of the transmission. This circuit works very well with most fm and a-m transceivers; however, some problems were experienced with certain models where the voltage on the unkeyed PTT line prevented the timing capacitor from discharging. This, of course, affects the duration of the resultant burst, or whether you get a burst at all. The problem has been eliminated in most cases by the addition of blocking diode D6 in the PTT line.

tone frequencies

The oscillator frequency is controlled by switching additional tuning capacitors across the 88-millihenery toroid. Available frequencies are limited only by switch size, room in the cabinet for tuning capacitors, and the low limit of the oscillator (approximately 1 Hz). If a progressively shorting switch is available, it will materially reduce the number of capacitors needed for multi-frequency operation. Adjustment for the various frequencies should begin with the highest frequency (lowest capacity) and progress to the lowest frequency (highest capacity).

transmitter control heads

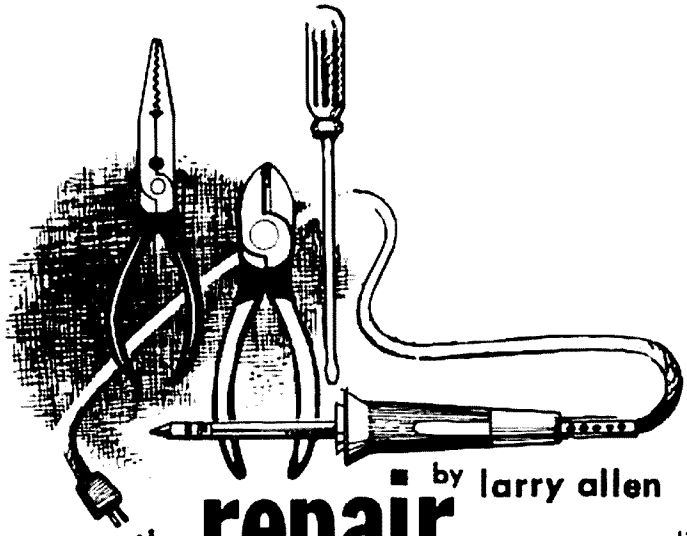
No changes are needed in the original transceiver equipment to accommodate this encoder. One change is required in the control head of each unit to provide power for the encoder. Most control heads use a 4-pin microphone connector. The fourth pin is for receiver audio to be used with a handset. This should be removed and the fourth pin connected to switched 12 volts dc in the control head. A 4-pole double-throw switch transfers the appropriate lines from one control head to the other.

Interconnection between encoder and control heads is via jumper cables made with standard microphone plugs to mate with the microphone jacks on the control heads. It is suggested that the transmitter deviation/audio level be set for each rig in the normal fashion, with the microphone plugged into the appropriate control head. Then the microphone should be removed and the jumper cables from the encoder connected. The individual levels in the encoder can then be adjusted to provide the same deviation as before. This technique allows instant restoration of facilities should disaster strike the encoder, or should you change rigs.

The tone level relative to speech peaks is controlled by the resistor in series with the dial pulser contacts. The value shown will normally provide tone deviation of the same magnitude as the speech peaks. This relationship can be varied as the system demands. The sine-wave tone level into a 100-ohm load is approximately 300 mV rms, usually enough to provide full deviation in most equipment. Carbon microphone excitation is provided by the preamp circuit, and the normal excitation from the set is blocked by the coupling capacitor.

In working with remotely located equipment, care must be exercised to ensure that no ground loops exist in the system. Note the audio common is isolated from chassis ground and is switched, as are the PTT and audio lines. The frequency stability of this oscillator is quite good. It remains within approximately 1 percent of the set frequency with wide load, voltage and temperature variations.

ham radio



by larry allen

the repair bench

troubleshooting amateur gear with an oscilloscope

Well, guess I'd better begin where I left off. Before my spring sabbatical, I promised to tell you about oscilloscopes. And so I will.

A scope lets you **look** at a radio or audio signal. That's why it's so valuable as a tool for troubleshooting. It measures signal voltages, and at the same time shows the shape of the waveform. Thus, if distortion in an amplifier has fouled up a signal so it doesn't come through plain enough to understand, a scope can show you which stage makes it distorted. In those transistor-switching power supplies, you can look at the switching waveform and even track down troubles **about** to happen. You're probably familiar with the scope as a modulation monitor. And, if you are interested in amateur television, you simply have to know how to use a scope.

Most hams don't bother with a scope much because they don't really know how. Hooking one up as a modulation monitor is simple compared to twisting the knobs so you can view some of the strange-looking waveforms a serious ham runs into.

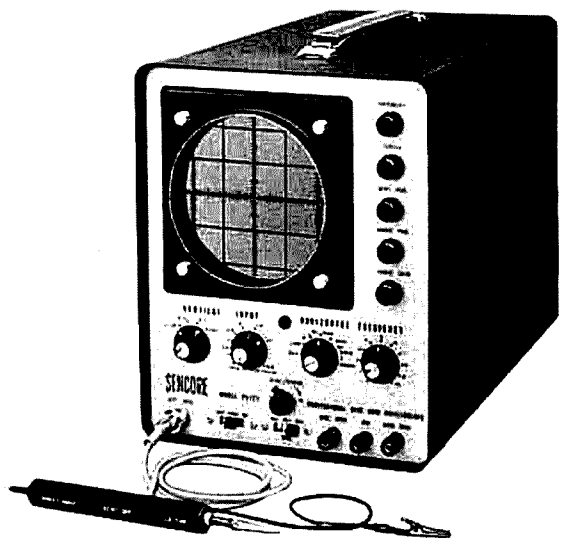
The scopes best suited for ham troubleshooting are the same kind used by radio and tv repair technicians. They should be wideband. You can buy them ready-made, like the one in **fig. 1.** Or, there are scope kits; **fig. 2** is one of those, in a compact size. The most common service-type models are recurrent-sweep scopes, called that to distinguish them from a more elaborate type called a triggered-sweep scope.

The dials and controls on the front of various brands and models differ only slightly. They're arranged in all sorts of positions, but the labels are always similar. Once you know one scope, you can quickly get acquainted with any of them.

getting the scope fired up

There's a group of knobs, usually beside or just below the screen, that turn the scope on, light up the trace, and get it set up to show a display. They are the **intensity, focus,**

fig. 1. One scope that's popular with tv servico technicians. Other well-known brands are listed in the box on page 57.



and **position** controls, and they're shown at the right in **fig. 3.**

The power switch may be on the **intensity** control. You turn it on, and turn the control all the way up. If the power switch is separate, turn the intensity up anyway. After a minute or so, either a dot or a line should show up on the screen.

Adjust the **focus** to make the dot or line as fine as possible. The top two photos of **fig. 3** show how a scope dot looks when it's out of focus and then when it's focused. The bottom two show out-of-focus and in-focus line traces.

If the dot or line doesn't show up after 1½ minutes, the two **position** knobs may be set wrong. Move them slowly from one end to the other, first separately and then simultaneously. **Vertical position** moves the display up and down on the screen, and **horizontal position** moves it from side to side. When the scope first comes on, if a **position** knob is near one end of its rotation, the dot or line may be out of sight. Set the knobs to

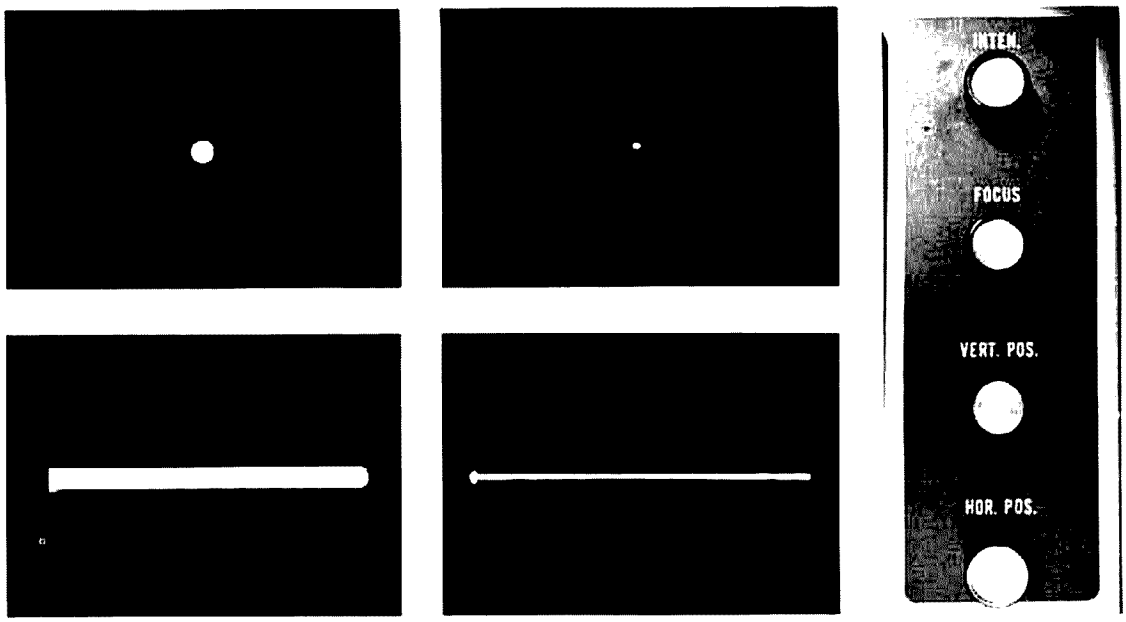


fig. 3. Scope dot or line when it's out-of-focus (large and fuzzy) and when it's in-focus (sharp and fine).

fig. 2. Compact, 3-inch scope you can build from a kit. Most kit-built scopes cost much less than comparable factory-built models.



center the line or dot as well as you can.

To make a dot into a line, which you'll have to do to display waveforms, find the group of controls shown in **fig. 4.** The important ones for this are **horizontal frequency** and **horizontal gain.** On some scopes, the main frequency dial is labeled **sweep.** It controls the sawtooth generator that scans the beam (which makes the dot) rapidly from side to side. Turn the **horizontal frequency** switch to a position that makes the sawtooth generator sweep the beam back and forth 20 or 30 times a second. The switch in **fig. 4** is set between the 10 and 100 marks. That's close enough for now; later, with a waveform on the screen, you can refine the frequency of the scope's internal sweep generator with the lower left control.

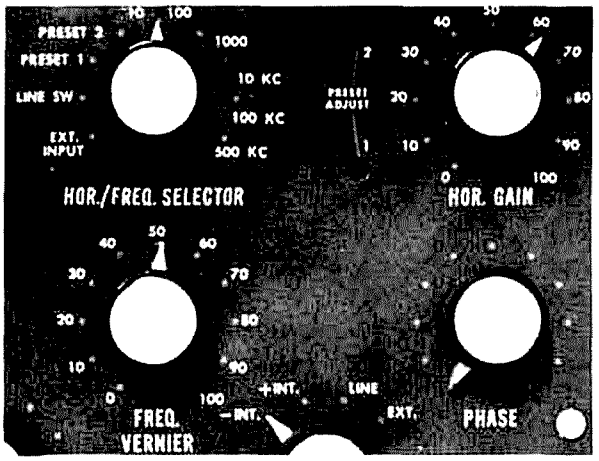


fig. 4. Froquency-control knobs on a kit-type servicing scope. Ignore phase knob at lower right.

There may still not be a line, or it may be very short. Turn up the **horizontal gain** until the line extends most of the way across the scope screen. The scope is ready now for you to display a waveform and learn how some of the other controls work.

looking at an ac waveform

One handy waveform is the 60-Hz voltage from the power line. In fact, so much radiation from it is picked up by the capacitance of your body, you can use that as a test signal. Just grab the tip of the probe with your fingers. Leave the ground clip dangling.

You'll have to set the **vertical input** controls (**fig. 5**) so the scope is sensitive enough to show the rather weak 60-Hz signal picked up by body capacitance. On the scope from which the **fig. 5** photo was taken, the main input knob is set to X1, the most sensitive position. As you can see from the photo, the **vertical gain** doesn't need to be turned up much. (It may on some scopes, especially if

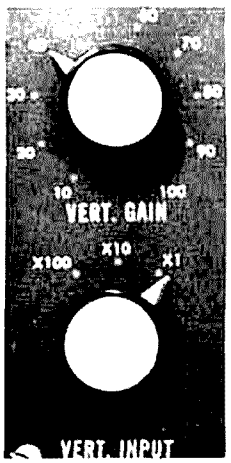
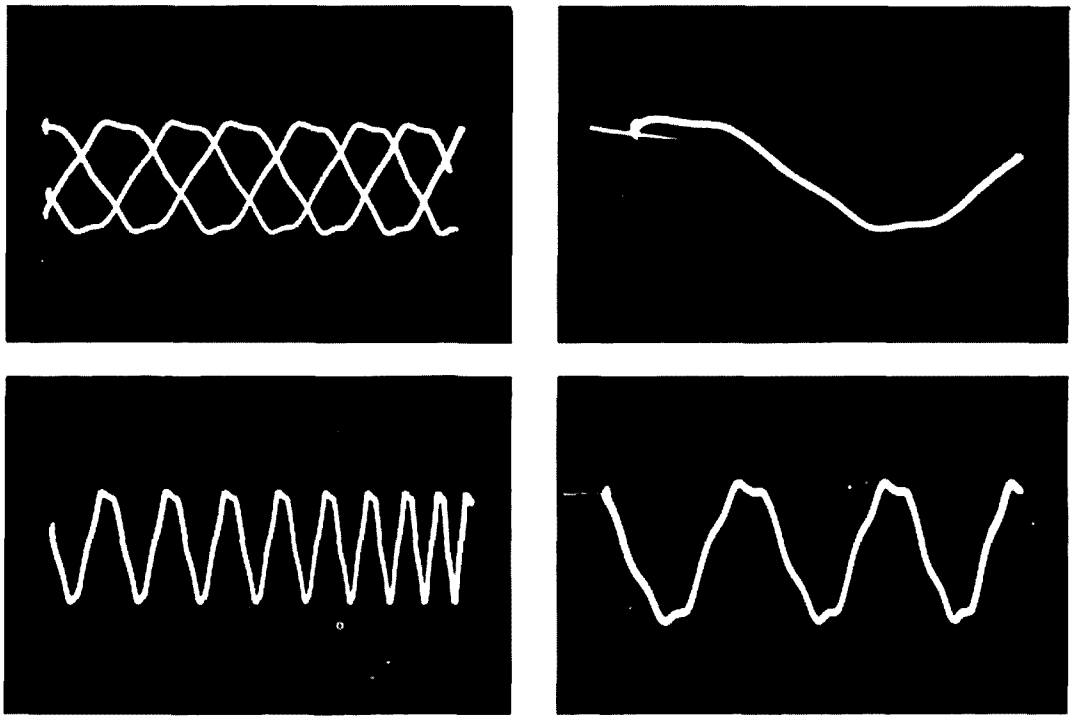


fig. 5. Vertical input knobs affect size of display on screen; on some scopes, they (along with graticule lines on the screen) help with voltage measurements.

fig. 6. Two top photos, and bottom left, show how the ac waveform may look before you adjust line frequency. Bottom right shows solidly locked three-full-cycle display.



60-Hz radiation is mild around your shack or bench.)

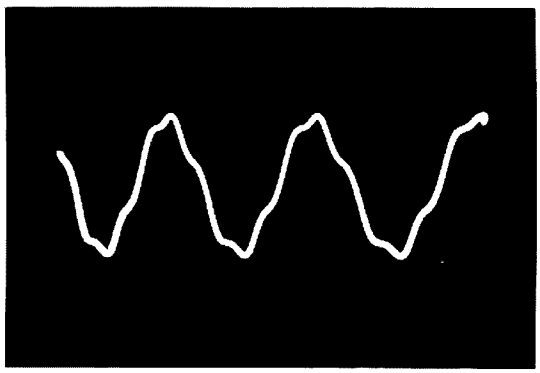
The waveform display you see on the screen, if you set the **horizontal frequency** switch as I mentioned earlier, looks something like the top photos or the bottom left one in **fig. 6.** Best viewing is with two or three cycles of a waveform on the trace. You have to adjust the fine frequency control (labeled **frequency vernier** on the scope in **fig. 4**). Just turn it carefully back and forth till you have only three waveforms on the screen—as in the bottom right photo in **fig. 6.**

The waveform may not snap into place quite that easily. It may appear to be running one way or the other, and you may not be able to stop it with the fine frequency control. Holding it in position solidly is the responsibility of a knob or switch labeled **sync.** In fact, you can adjust where the left side of the trace starts, merely by how you set **sync.**

In scopes that have only a **sync** switch, you have a choice of positive-going (+) or negative-going (—) internal sync, external sync, or sync from the scope's power transformer (line). You'll almost always use internal sync, and whether you choose + or — depends on the general shape of the waveform.

With the waveform at top left in **fig. 7,** the signal display is shown synchronized on the negative-going portion of the test waveform; the **sync** switch is at **—int.** In the top right photo, the waveform is synchronized on the positive-going part of the signal; the **sync** switch is set at **+int.**

A scope with a control instead of a switch is synchronized much the same. The knob is usually marked with a center zero, and can be turned either way. One direction locks the signal on the negative slope, and the other locks it on the positive slope. If your scope has the control, notice once you get the display stopped how you can work the control back and forth slightly and shift the point on the slope where the trace starts. That isn't too important with an ordinary sine-wave display, but it is when you work with oddly shaped waveforms, like in amateur tv.



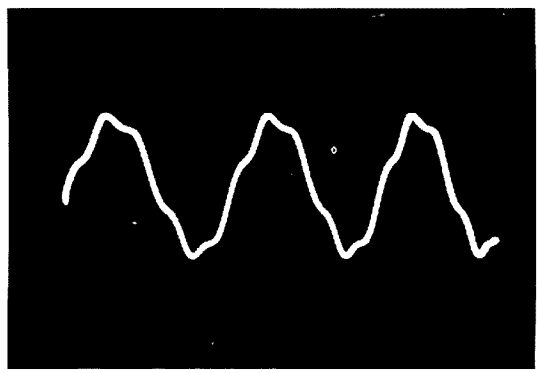
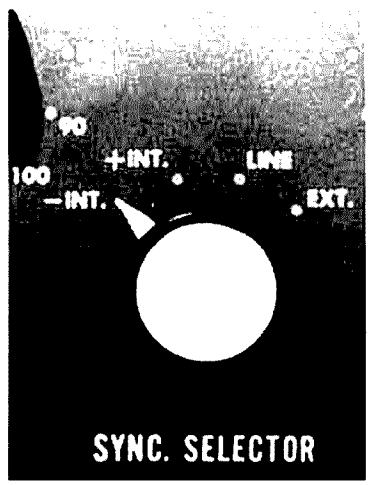


fig. 7. Ac waveforms locked, by sync switch, on negative slope (left) and positive slope (right). Irregular shape of sine ac is due to fluorescent-lamp radiation.



looking at odd waveshapes

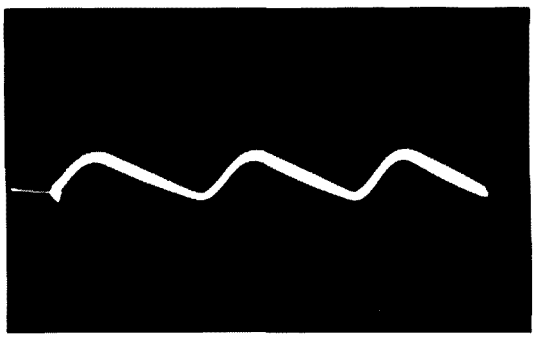
Sine waves are the most common type you'll look at when you use the scope on your repair bench. They're what you find in the power supply before the rectifier and filters get at the voltage. And you often in-

ject sine waves and then trace them to check out audio amplifiers, clippers and other speech stages.

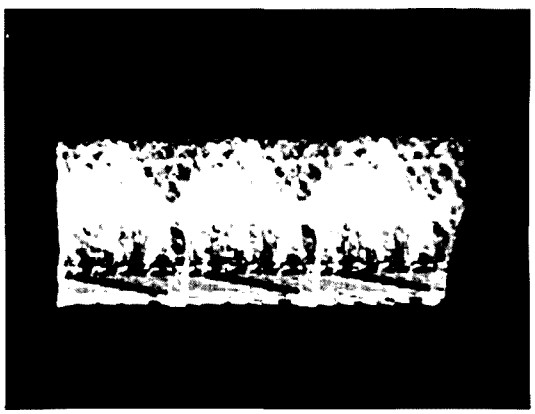
But sine waves are far from being the only waveshape you'll see. The only problem you can have viewing the odd-shaped ones is making them stand still on the screen. But there are a couple of simple tricks for that. If you know what's important about a waveform, that's enough to let you sync it in tightly on the scope—every time.

As one example, look at the rounded sawtooth at the top in **fig. 8.** It's taken at the

fig. 8. To lock in unusual waveshapes, you need to know their approximate frequency and which direction the dominant slope goes. All of these were taken in a home-tv flying-spot scanner.







output of a rectifier, across the input filter. The important thing to notice is the steepest slope. That's always easiest to synchronize on. In this one, the steepest slope is upward, which means it goes in the positive direction. So, you set the **sync** switch to the **+** position. And, as you can see in the photo, the trace locks (and starts) on the upward slope.

You have to learn to recognize the dominant steep slope and which way it goes. Often the waveform can only be slowed down a little with the **frequency** control, and you have to see the waveshape "on the run." Once you're sure which way the steep slope goes, you can set the sync for the polarity that will lock easiest. Then you can go back to the fine frequency control and slow the waveform down from running. When you get the scope sweep close enough to an exact multiple of the frequency of the signal you're viewing, the waveform locks in.

An example of locking in an odd-shaped waveform appears at the lower left in **fig. 8.** This is the video waveform from a flying-spot scanner such as you might build for ham tv. This display shows three cycles of video, which means the scope's sweep must be set to one-third the horizontal scanning rate of the tv set.

The horizontal sweep for commercial TV is 15,750 Hz. The scope, then, must be set to about 5,250 Hz. That way, the scope display has time to show three cycles during each scan of the scope beam. The **horizontal frequency** switch in **fig. 4** must be set between the **1000** and **10kc** (10 kHz) marks. Then the **frequency vernier** is turned slowly back and forth until you catch a glimpse of the shape of the waveform.

You'll see that its dominant slope is downward-going, meaning the waveshape is main-

ly negative. So, to lock it in solid, the **sync** switch is set on the **—int** mark.

With the sync switch inadvertently set wrong, the waveform can't be settled down much better than in the lower right picture in **fig. 8.** This photo was taken with the switch at **+int.** You can see only a vague outline of the waveshape, if you look close.

using a scope for troubleshooting

Until you've used an oscilloscope for tracing faults in ham units, you can't appreciate how helpful it can be. And it's a mistake to try working on ham tv without one. Now that you know how to stabilize the waveforms on the screen, you can put a scope to work on your repair bench.

Next month, I'll tell you more about it. There are dozens of different tests you can make with it, and lots of ways to save time. I'll show you the waveforms you can expect to find in different kinds of equipment, how to measure them, and how to tell if they're not what they should be. And I'll tell you how to set the scope to look at each of them.

oscilloscope manufacturers

B&K Division, Dynascan Corporation, 1801 West Belle Plaine Avenue, Chicago, Illinois 60613

EICO Electrical Instrument Company, Inc., 283 Malta Street, Brooklyn, New York 11207

Heath Company, Benton Harbor, Michigan 49023

Hickok Electrical Instrument Company, Inc., 10523 DuPont Avenue, Cleveland, Ohio 44108

Jackson Electrical Instruments, 124 McDonough Street, Dayton, Ohio 45402

Knight Electronics Company, 100 North Western Avenue, Chicago, Illinois 60680

Leader Instruments Corporation, 24-20 Jackson Avenue, Long Island City, New York 11101

RCA Parts and Accessories, Deptford, New Jersey 08096

Sencore, Inc., 426 South Westgate Drive, Addison, Illinois 60101

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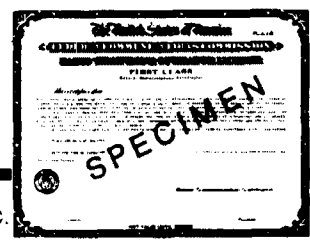
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ionospheric propagation

Victor R. Frank, WB6KAP, 12450 Skyline Boulevard, Woodside, California 94062

This month we will investigate ionospheric propagation quantitatively using some of the most powerful aids available to the ionospheric physicist. The results will be far from general, but I hope they will be enlightening. We will need a model: a profile of ionospheric electron density vs height. We could get such a profile by direct measurement with a rocket probe but measurements can be made from earth-bound ionospheric sounders more conveniently and inexpensively.

The ionosonde is an upward-looking pulsed high-frequency radar whose frequency is swept in a predetermined manner (logarithmically on most ionosondes presently in use). Echoes are obtained from the overhead ionosphere, and from the variation of time delay between transmitted and received pulses vs frequency, a profile of electron density vs height can be determined.

A variety of echoes may be present. If we restrict our attention to those that involve only one passage from earth to ionosphere and return, there are, in general, two echoes returned at any given frequency below that which allows penetration of the ionosphere. The two echoes are from the ordinary and extraordinary waves. In order to explain the reflection process at vertical incidence I will describe the interaction between a radio wave and the free electrons in the ionosphere.

properties of ionized media

High-frequency radio waves are propagated to great distances by virtue of the presence of free electrons in the ionosphere. The neutral particle density is small enough that collisions between particles are relatively infrequent and electrons are free to move in response to electric and magnetic fields. In addition to random thermal motions and large-scale drifts, electrons have orderly oscillatory motions in response to radio frequency electromagnetic waves. The contributions of the positive ions may be neglected at high frequencies since these ions are much more massive than the electrons, and they do not acquire any significant velocity or displacement in response to the high-frequency fields.

A 1 kW transmitter at 14 MHz fed into a three element Yagi might produce the following electron motion in the E region:[1]

maximum displacement	4×10^{-6} cm
maximum velocity	350 cm/sec
number of cycles between collisions	140

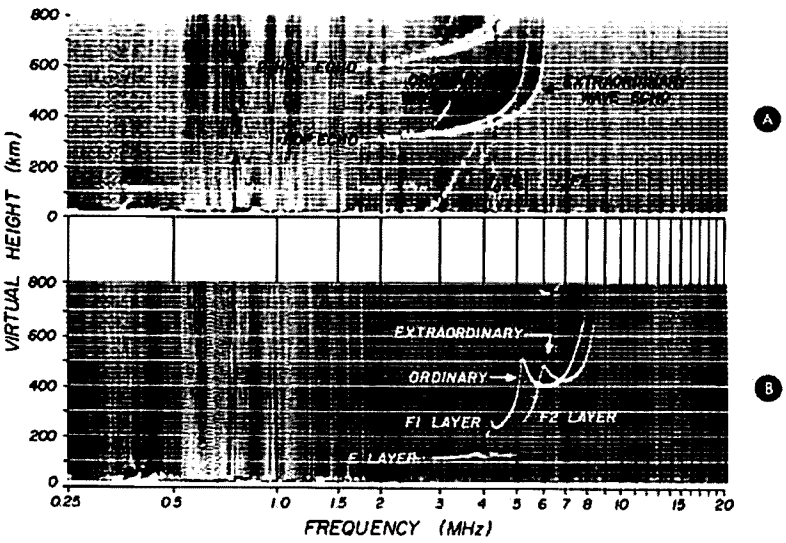
For comparison, at 300° K (room temperature) in the E region, the displacement and velocity due to thermal motions are:

mean free path	100 cm
thermal velocity	10^7 cm/sec

The motion of an electron is a current flow, and a scattered field is radiated from each electron. The combined effect of trillions of free electrons moving in concert is the formation of weak scattered fields which are approximately 90° out of phase with the original field. The vector sum of the scattered and incident waves results in amplitude almost equal to the original field but with an advance in phase. The scattered fields add constructively only in the direction of progagation, while creating a large total advance in the phase of the electromagnetic wave relative to that which would occur in a vacuum.

We describe this modification of the wave by assigning an index of refraction to the medium. The behavior in an ionized medium is identical to that which would be expected in a dielectric whose index of refraction is less than unity. (Note that for dielectrics with

fig. 1. Ionograms taken by vertical-incidence sounder at Pt. Arguello, California (34.5° N. latitude) on August 14, 1968 at 2200 pst (A) and August 15, 1968 at 1100 pst (B).



bound electrons, the resultant is a smaller decrease in phase and an index of refraction greater than unity). The index of refraction of the ionosphere depends not only on the electron density, but also on the frequency of the electromagnetic wave. Thus the ionosphere is a dispersive medium.

The index of refraction, μ, for the ordinary wave is

$$\mu = \sqrt{1 - \frac{80.5N}{f^2}} = \sqrt{1 - \frac{fN^2}{f^2}}$$

where N is the electron density per cubic centimeter and f is the radio frequency in kilohertz. At a radio frequency equal to the plasma frequency, fN, the index of refraction is zero for the ordinary wave. Propagation to regions of higher electron density is not possible and reflection occurs.

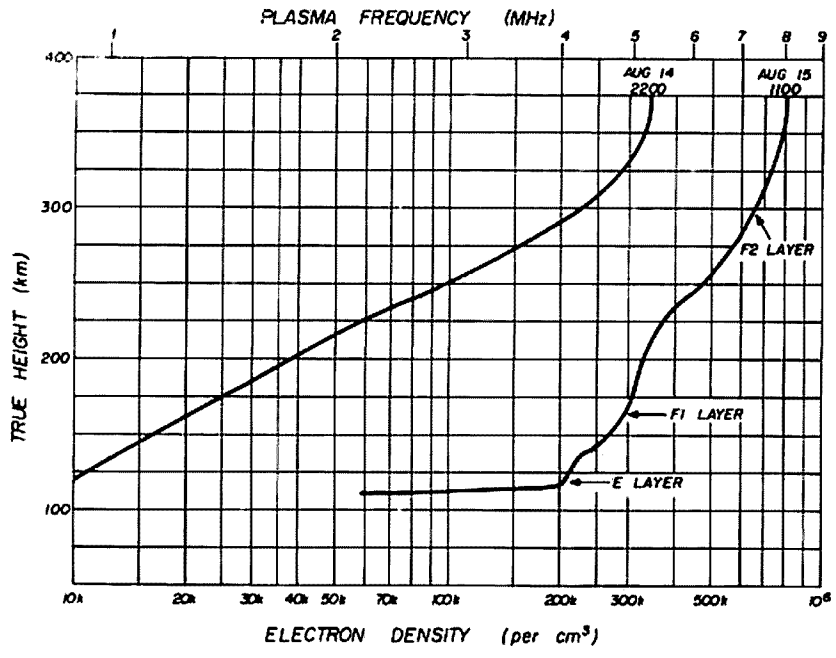
Fig. 1 shows two ionograms taken at Point Arguello, California on August 14, 1968 at 2200 pst and on August 15, 1968 at 1100 pst. The ionograms show traces of virtual height vs radio frequency from 250 kHz to 20 MHz. The virtual height is the time delay times the free-space velocity of light. The true height of reflection is less than the virtual height since the wave is retarded by its interaction with the electrons in the ionosphere. The group velocity, v_G, of the ordinary wave is approximately

$$v_G \cong \frac{c^2}{v_P} \cong \frac{c}{\mu}$$

where c is the velocity of light in a vacuum, and v_P is the phase velocity. Thus the wave is greatly retarded in regions where the refractive index is small over a wide range of heights (near the heights of maximum ionization).

There are preferred directions of electron motion due to the presence of the earth's magnetic field. This anisotropy* results in multiple values of the refractive index and, in

fig. 2. True height profiles of electron density vs height derived from the ionograms of fig. 1.



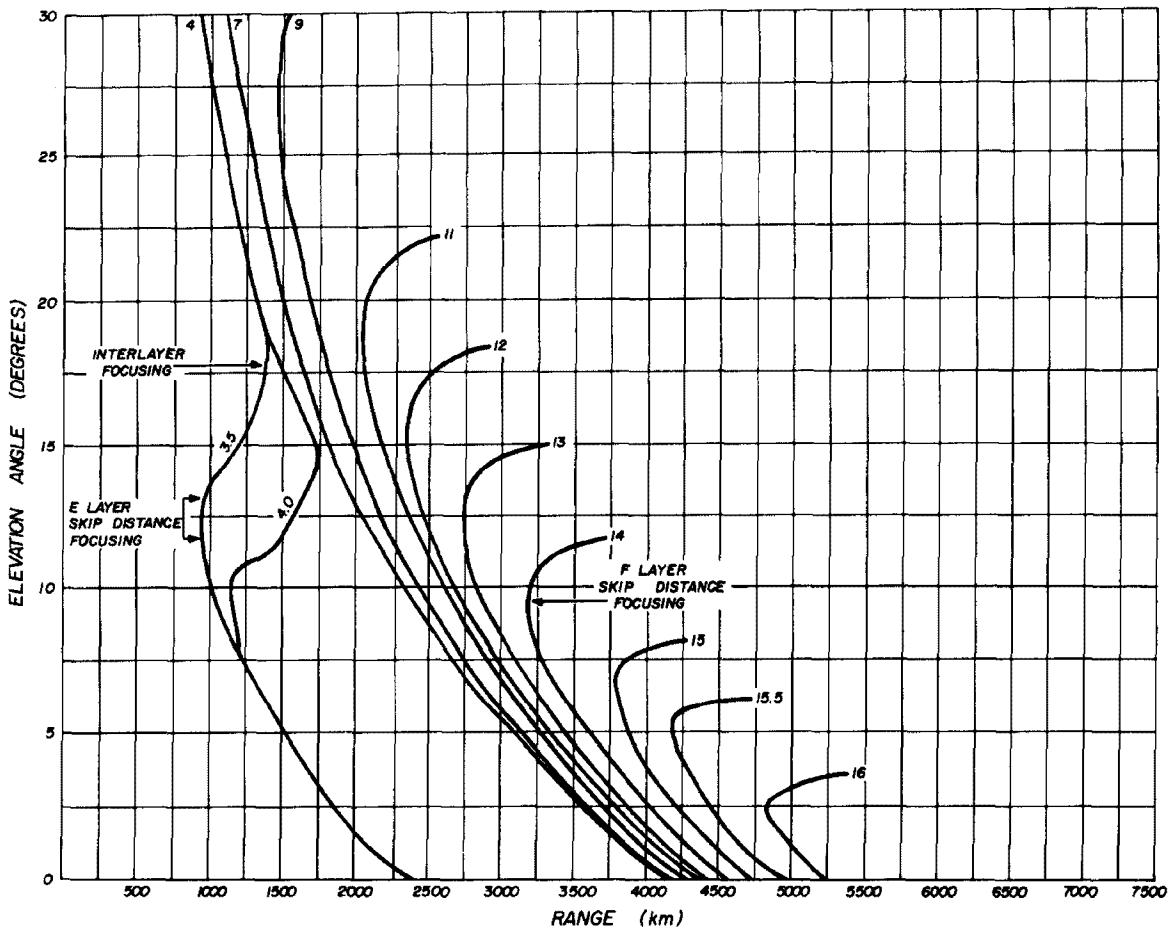


fig. 3. Range vs elevation angle plots for the ionogram of fig. 1A (2200 hours).

general, separation of a linear polarized wave into two counter-rotating elliptically-polarized characteristic waves which may travel over independent paths in the ionosphere. These characteristic waves are called the ordinary and extraordinary waves. Although the extraordinary-wave muf is higher than the ordinary-wave muf, it is dependent on the direction of propagation with respect to the magnetic field, and thus is more difficult to calculate. All of the calculations in this column are for the ordinary wave only.

There are approximate graphical aids for determining reflection heights, skip distances and penetration angles for oblique propagation from vertical-incidence ionograms. Ionogram scalings reported in past columns have used these transmission curves. For further information on transmission curves, you may want to consult references 2 and 3. This month I will show a more accurate (and expensive) technique that may be used if you have access to a digital computer.

First a model of the ionosphere is prepared from a vertical-incidence sounding (assumed to be taken at the path midpoint). A true height profile of electron density vs height is calculated.[4] The true height profiles (**fig. 2**) are, in general, not unique. They assume a monotonic increase of electron density with height; occasionally there are "valleys" between the E and F1 layers that are not directly measureable with ionospheric sounders. Another problem in creating models of the ionosphere is that the ionogram traces are seldom continuous. Absorption, interference and poor antennas frequently limit the low frequency coverage. However, the true height profiles obtained are probably as useful as any for predicting what the ionosphere may be like at mid-latitude United States near 1100 and 2200 local time during August 1969.

* Anisotropy describes a medium that exhibits different properties when measured along axis in different directions.

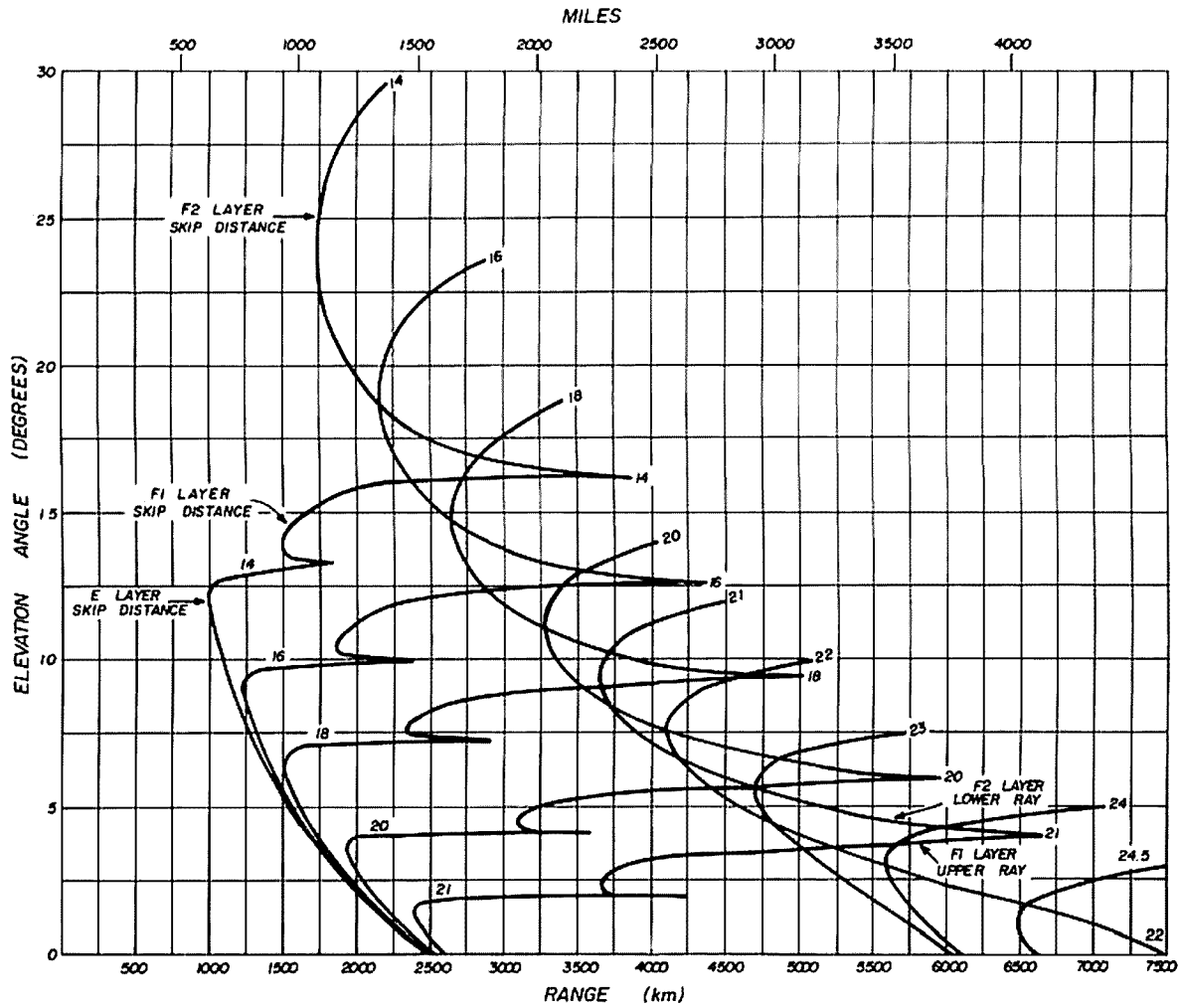


fig. 4. Range vs elevation angle plots for the ionogram of fig. 1B (1100 hours).

Given an ionospheric model, you can trace the path of a ray launched at a given angle at a given frequency. Using a program developed by Dr. T. A. Croft at Stanford. The following equation is solved in many steps

$$\mu r \sin \emptyset = K$$

where r is the distance from the center of the earth, $\emptyset$ is the angle of incidence, and K is a constant for any given ray called the characteristic. (This is Bouger's rule, the equivalent of Snell's Law for circular coordinates). These calculations produce propagation time delays, ground range and reflection height for rays of a given elevation angle at any particular frequency. For this column I have chosen to plot curves of ground range of single-hop propagation vs elevation angle for various radio frequencies. These curves are shown in figs. 3 and 4 for 2200 and 1100 local time, respectively.

Note from fig. 3 that at a given frequency

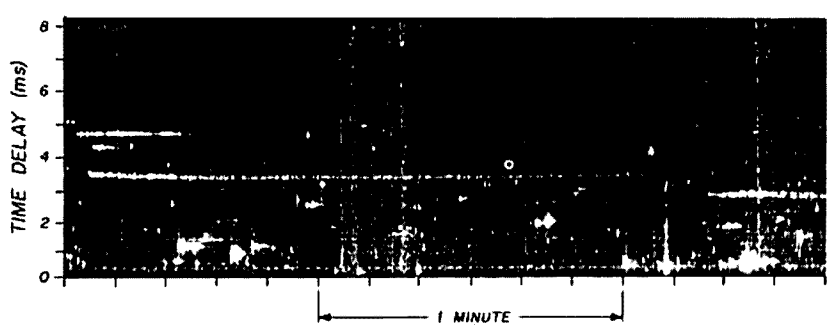


fig. 5. Range (delay time) vs time display of backscatter meteor echoes on a frequency of 25 MHz during last year's Perseids meteor shower.

two rays may be returned to the same distance. The ray with the lowest elevation angle for a particular layer, called the lower ray, is usually strongest and occurs over the widest range of distances. The upper (or Pederson) ray is most frequently observed at frequencies just under the muf for that distance. For some ionospheric models and frequencies, the upper ray may be received at greater distances than the lower ray. If antennas with poor low-angle response are used and the distance is greater than 2500 miles or so, the maximum observed frequency may be propagated by single-hop upper rays.

Note from **fig. 4** that the situation is compounded during daylight hours by the possible presence of E- and F1-layer modes as well. The extraordinary waves (not computed) will almost match the ordinary waves, with the greatest separations at short distances and near the skip distances.

Curves of ground range vs elevation angle point up the focussing and defocussing that occur in the inosphere. Focussing occurs at the skip distance where neighboring rays arrive very close to the same distance. The upper rays are defocussed at frequencies much below the muf (or distances much further than the skip distance) since neighboring rays are spread over a wide distance. The skip distance for frequencies near the absolute muf is much greater than the 2500 miles usually assumed for the maximum distance for one F2-layer hop. When considerable ionization is present below the F2-layer, the maximum one-hop range may occur for elevation angles other than zero or just below the penetration angle (see curve for 21 MHz, 1100 hours).

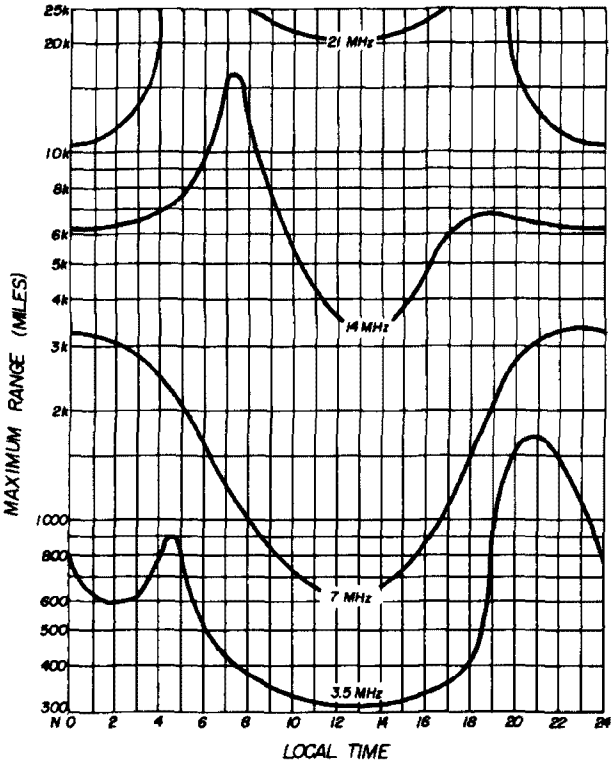


fig. 7. Maximum range to the north from 38° N. latitude as limited by absorption, atmospheric noise and system parameters (100 watts cw to typical antennas).

However, predictions of extreme range upper rays should be viewed with some pessi-

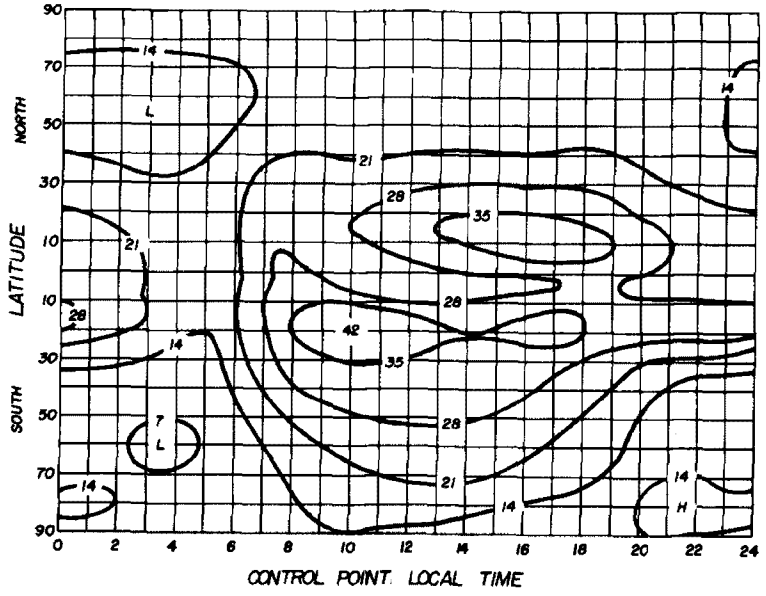


fig. 6. Time chart of muf-(4000)F2, the median maximum usable frequency for a 2500-mile path predicted by ESSA for August 1969 based on 120° W. longitude.

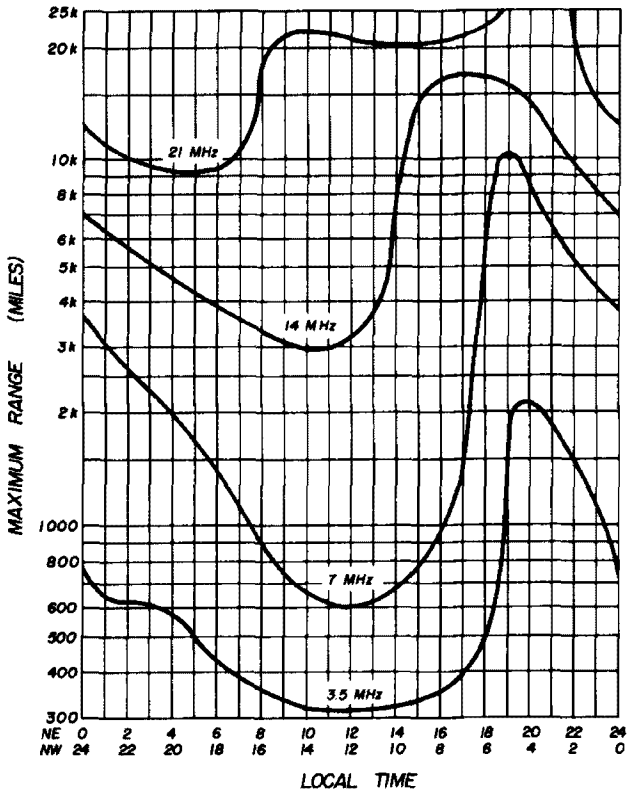


fig. 8. Maximum range to the northeast (top time scale) and the northwest (bottom time scale) from 38° N. latitude.

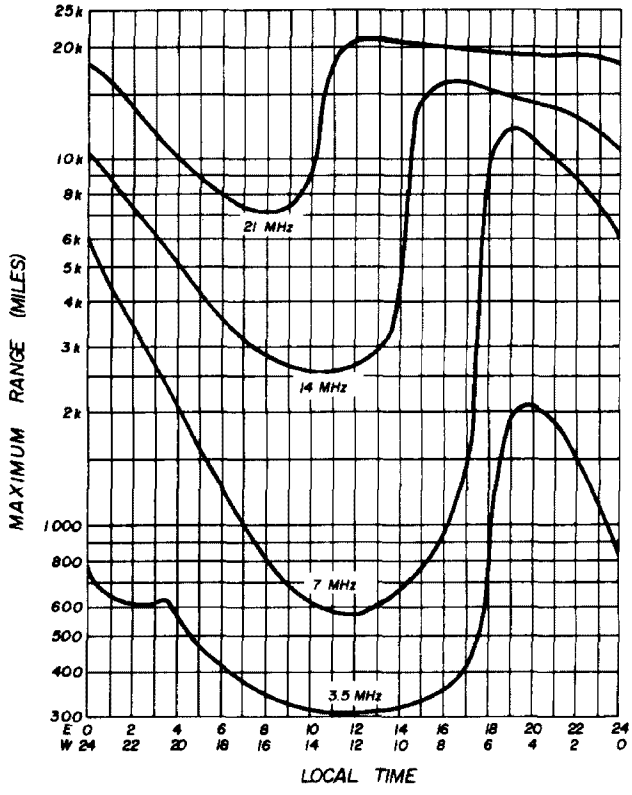


fig. 9. Maximum range to the east (top time scale) and to the west (bottom time scale) from 38° N. latitude.

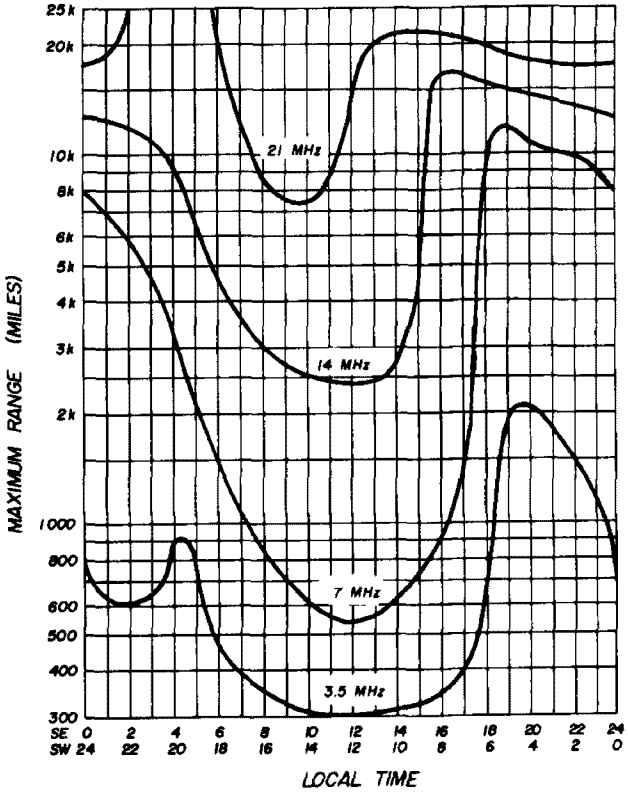


fig. 10. Maximum range to the southeast (top time scale) and the southwest (botom time scale) from 38° N. latitude.

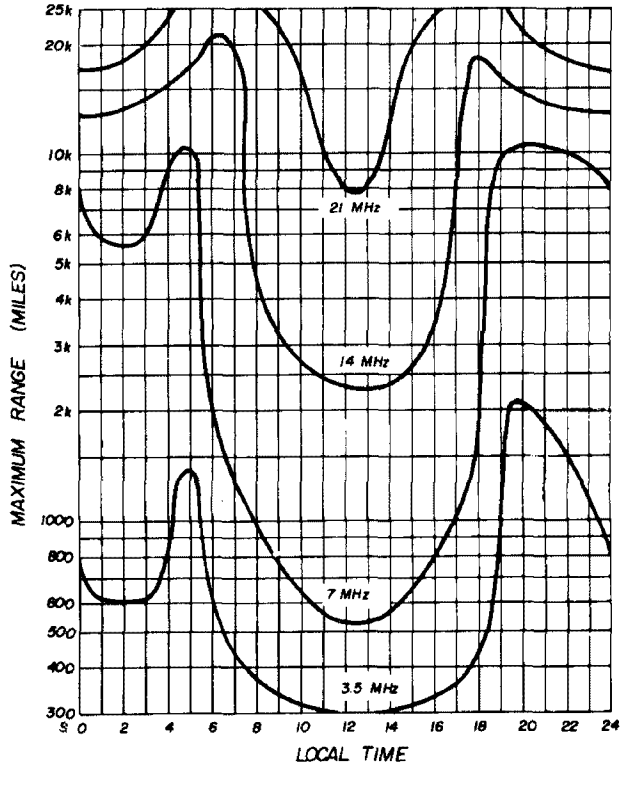


fig. 11. Maximum range to the south from 38° N. latitude.

mism as the ionospheric model was assumed to not vary with distance. This is a doubtful assumption for rays that remain in the ionosphere over such distances. Usually the rays will be returned to earth at shorter distances if the electron densities increase with distance —or will either penetrate the ionosphere or be trapped in ionosphere-ionosphere modes if the electron densities decrease with distance.

The exact characteristics of the transition region between modes depends critically on the shape of the ionosphere profile. When the layers are not well defined, the transition may occur as a turn instead of a cusp (3.5 and 4 MHz, **fig. 3**). In this case, another region of focussing occurs at the maximum of range for a single hop by the F2-layer.[6,7]

meteors

August is a good month for meteors. The Perseids meteor shower (August 10 to 14) is responsible for many 144-MHz contacts over distances of about 600 to 1400 miles each year. Meteors by themselves are of little interest to the radio amateur; of interest is the ionized trail which results when the meteor enters the earth's atmosphere and is vaporized. Backscatter from meteor trails is also noticed at 50 MHz and lower frequencies.

Fig. 5 is a record of reflections from many meteor trails observed near 25 MHz during last year's perseids meteor shower. The range to a meteor trail is about 150 km (93 miles) per millisecond delay. Some meteor trails were observed as far away as 800 km, but most were between 100 km and 600 km. Note the trail at 540 km (3.6 ms) that lasted two minutes. More frequent, of course, are the bursts of a second or less which are of limited use for ordinary radio amateur low-speed transmission.

Other meteor shows in August include the Cygnids (August 10 to August 20) and the Draconids (August 21 to August 31).

vhf propagation during may

Reports of long distance vhf propagation during the first half of May are down from April, but lots of DX was worked on 50 MHz. As of May 20, ZK1AA had over 50 contacts in less than 2 months, all in the Northern Hemisphere and over 2900 miles. His beacon is now operating daily from 1730-0930 gmt. Trevor, 5W1AR, now has a beacon on 50.105 MHz operating 1900-2400 gmt and 0400-0600 gmt. Last I heard he was having tvi problems with tv from American Samoa which comes on at 0600 gmt. A chronological list of TE (transequatorial) contacts follows:

May 4, 0330-0500 gmt, ZK1AA worked WB6YPF, WA6HXW, K6QEH, W6RUX, WA6GJU and W6NIT

May 6, ZK1AA worked KH6GRU; 2245 gmt, W4GDS and WA4MHS

May 11, 0650 gmt, ZK1AA worked WB6KAP; 0900 gmt, ZK1AA worked KH6GRU; KH6GRU heard some VK's

May 18, 0630 gmt, ZK1AA worked JA1AC and heard a Japanese broadcast station on 56 MHz; 0730 gmt, ZK1AA heard KH6 fm stations to 95 MHz.

May 28, 0738-0823 gmt, WB6KAP heard ZK1AA beacon

Note a shift in the times of TE occurrences as summer approached. The ZK1-to-KH6 path has been opening as late as 0700-1100 gmt, and on May 28 this path was open only from 1000-1030 gmt; these times are after midnight on the West Coast!

propagation for august

Shortwave conditions this August should be very similar to those of last year. Comments made in last August's column are equally applicable. Summarizing, some improvement is expected over July in higher daytime muf's, lower absorption and lower noise levels. Sporadic-E activity is expected to decline while transequatorial openings are expected to reappear to the Southern States. There is the possibility of vhf aurora openings (or disturbances—depends on the band you're interested in) to the northern states.

The F2-layer muf chart for this month (**fig. 6**) is based on 120° W. longitude. The muf chart that appeared last year (for 75° W. longitude) is also still valid.

Note that all time scales are local solar time, which is within 30 minutes of local **standard**

how to use these propagation charts

1. To find the maximum usable frequency for F2-layer propagation for distances of 2500 miles or more in any direction, find your control point and read the frequency from the F2-layer muf time chart. Your F2-layer control point is 1200 mies away from your station in the direction of propagation; this is about an 18-degree difference in latitude for a north-south path, or 1½ hours time difference for an east-west path.

2. To determine the path muf for a path under 2500 miles, the control point is at the path midpoint, and a correction factor is applied to the 2500 mile muf. These correction factors are plotted in fig. 7 of the August 1968 column.

3. During summer daytime, the path muf for a path shorter than 1200 miles may be set by E-layer propagation. To determine if this is so, refer to the E-layer muf time chart and the chart of muf reduction factor vs distance in this month's column. Note, however, that sporadic-E will probably result in zero skip distance on 7 MHz and skip distances under 400 miles on 14 MHz very frequently during daylight and evening hours.

4. The F2-layer will probably be effectively shielded by the E-layer for operating frequencies below 70 percent of the predicted E-layer muf.

5. Over any particular path involving more than one hop, the path muf is the lower of yours and the other station's control-point muf. The muf time charts may be treated as muf contour maps. (The F2-layer chart has significant errors outside the range of longitudes between 45°W. and 135°W.) As such, each hour is the equivalent of 15° of longitude. A map drawn to the same scale can be overlayed and positioned to the right or left to show the variation of the muf contour map with time. Curved lines may be drawn on the overlay representing great circle paths, as found from a globe or "Ionospheric Radio Propagation," printed by the U. S. Government Printing Office.

6. To find the maximum propagation distance as limited by ionospheric absorption and atmospheric noise, refer to the maximum range charts for the directions you wish to work. Note that the time scales are reversed for westward propagation. Also note that this month the noise curves have not been assumed to be symmetrical about local noon. Thus when the curves are used to predict propagation to the west they may be in error due to this lack of symmetry. These curves are based on a unity signal-to-noise ratio of a 6-kHz bandwidth with 100 watts cutput and antenna gains (over an isotrope) of 6 dB for 20 and 15 meters, 0 dB for 40 meters, and —6 dB for 80 meters at each station.

time, not daylight savings time. Also note that the muf contours of interest are those for the time and latitude of your control points, 1200 miles away in the direction of propagation.

next month

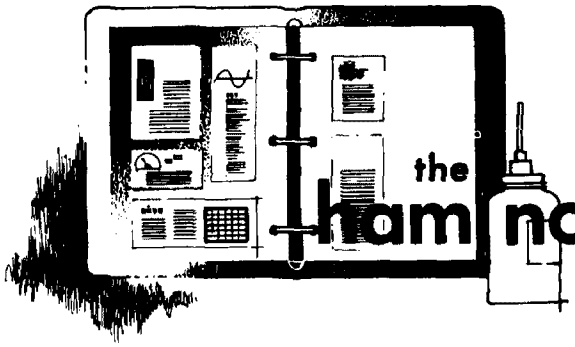
Beginning with the September issue, Propagation predictions will no longer be included as a part of **ham radio**. Readers who are interested in propagation information may obtain a monthly propagation bulletin via first class mail by simply writing to the editor. This bulletin will provide you with the latest up-to-the-minute propagation forecasts.

Note: The formula for the maximum electron density of the E-region given in the May 1969 column on page 62 is incomplete. It should have been:

$$N = 10^4 (180 + 1.44R) \cos^{1/2} \chi$$

references

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2. NBS Circular 462, "Ionospheric Radio Propagation," U. S. Government Printing Office, Washington, D. C., 1948.
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4. K. G. Budden, "Radio Waves in the Ionosphere," Cambridge University Press, 1961.
5. T. A. Croft, "Methods and Applications of Computer Raytracing," Technical Report No. 112, Contract NONR 225(64), Radioscience Laboratory, Stanford University, January, 1969.
6. T. A. Croft, "HF Radio Focusing Caused By the Electron Distribution Between Ionospheric Layers," *Journal of Geophysical Research*, 72, May, 1967, p. 2343.
7. T. A. Croft and T. W. Washburn, "Comparison of Simultaneous Forward and Backscatter Soundings to Verify a Proposed Focusing Mechanism of HF Sky Waves," *Journal of Geophysical Research*, May, 1969, p. 2443.

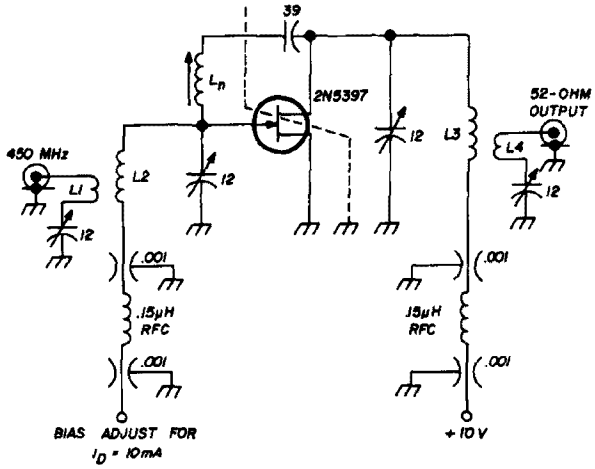


fet preamp for 432 MHz

If you're trying to get the most out of your receiving gear for 432 MHz, you should look into the new 2N5397 and 2N5398 field-effect transistors manufactured by Siliconix.* These new devices are usable up to 800 MHz and their performance on 432 is nothing short of fantastic. Manufacturer's claims are many times on the optimistic side, but according to K6JYO and K6KV, the 2N5397 performs as advertised—at 450 MHz the noise figure is 2.5 dB typical (3.5 maximum), and power gain in the manufacturer's 450-MHz circuit in **fig. 1** is 18 dB. The 2N5397 is $8.50 and the 2N5398 is $6.00.

*Siliconix Incorporated, 1140 West Evelyn Avenue, Sunnyvale, California 94088

In the common-source amplifier shown in **fig. 1**, the drain and gate leads should be shielded from each other to reduce feedback phase shift. The neutralizing coil should be low loss for best gain and noise figure. One other construction tip: tune patiently!



L1 1.4" long piece of no. 22 enamelled, spaced 0.1" from L2

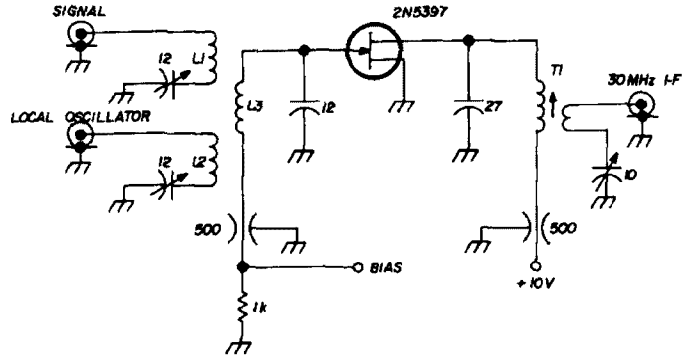
L2 1.1" long piece of no. 16 solid copper

L3 1.3" long piece of no. 16 solid copper

L4 1.4" long piece of no. 22 enamelled, spaced 0.3" from L3

Ln 3 turns no. 22 enamelled on ¼" slug-tuned ceramic form, aluminum slug

fig. 1. 450-MHz common-source amplifier provides 18-dB gain.



L1 1.4" long piece of no. 22 enamelled, close coupled to L3

L2 1.6" long piece of no. 22 enamelled, close coupled to L3

L3 1.75" long piece of no. 18 copper

T1 primary 13 turns no. 22 enamelled, close wound on ¼" slug-tuned form; secondary is 3 turns no. 22 enamelled, close wound over primary

fig. 2. 450-MHz common-source mixer circuit.

Another circuit for the 2N5397 is shown in **fig. 2**; this is a common-source mixer. When building this circuit, bypass the drain lead as close to the transistor case as possible. Maintain a high impedance at the intermediate frequency for maximum gain and maintain a signal ground on the source. Tune the tank circuit for best gain and noise figure.

Jim Fisk, W1DTY

using Swan 350 and 400 equipment for rtty

I've received many inquiries from owners of Swan 350 and 400 transmitters on how to put this equipment to work in the rtty mode. Much correspondence resulted in the methods described here.

Once you've acquired the teletypewriter, you can add simple circuits for rtty operation. Two methods, afsk and fsk, were tried by WB2UCI and me with satisfactory results.

The afsk method involves feeding a sine-wave audio signal into the speech amplifier via the microphone jack. This shifts the modulating frequency, which produces frequency shift keying. In the fsk method, a keyer (circuit) inserts a small amount of capacitance into the vfo to vary its resonant frequency at the keyed rate.

The transmitter carrier is changed at an 850-Hz rate. This consists of two frequencies 850 Hz apart, called the space and mark frequencies. The standard frequencies for these are 2125 Hz for mark and 2975 Hz for space. The Swan 400 and 350, however, have an audio cutoff at 2450 Hz, so they won't pass 2975 Hz. The remedy for this is to change the space and mark frequencies to 2125 Hz and 1275 Hz, whose difference is still 850 Hz.

Essco teletype specialists offer the TU-7 unit which is already modified for the Swan transceivers at moderate cost. It's battery operated and also serves as a demodulator for receiving. Further information may be acquired by writing to Mr. J. S. Tessler.*

The fsk method uses a keyer to add a small amount of capacitance into the emitter of the vfo oscillator transistor. The circuit that seems to work best with the Swan is shown in **fig. 3.** (The complete circuit appeared in the May 1965 issue of **QST.**) The keyer circuit is constructed on a three-lug terminal strip and is mounted inside the vfo. The lead from the trimmer capacitor to the emitter of the transistor should be as short as possible. Care must be taken when wiring this lead, since the transistor's lead is extremely short, and the printed circuit board is very difficult to remove. The methods described work very well with both Swan units.

* Essco, 324 Arch Street, Camden, New Jersey.

Outboard vfo's would make rtty much more enjoyable and less tedious. When using the transceiver alone, the transmitting and receiving frequency will not be the same, which may present a few problems. With another vfo the receiving and transmitting frequencies will be independent of each other.

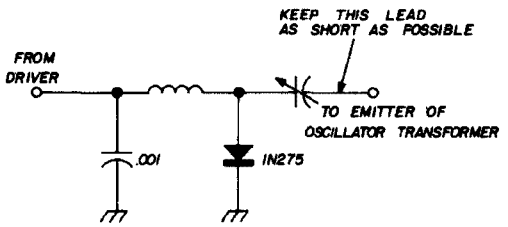


fig. 3. Circuit for fsk method in Swan 350 and 400. Mount it on a 3-lug tie point and install inside vfo section.

Tuning the Swan with these two methods is basically the same as in regular operation. When using afsk, tune the transmitter exactly as for ssb, making sure there's no carrier or other sideband present. If this occurs, afsk will be generated, which is illegal on the low bands.

With the keyer circuit, load the same as for a-m and turn off the microphone gain. The Swan will now operate in F1 mode at 120 watts. This is adequate for cross-country contacts and occasional DX when the band opens.

Joseph Boniakowski, WB2MIC

the indispensible glass

As age creeps on, one of the handiest gadgets in the ham shack is the old reliable magnifying glass. For example, if a DX station tells you to move up 2 kHz, and you find that your glasses have been left upstairs, the magnifying glass will let you easily read the spaces on your dial.

At the same time, a humorist will tell you that it's fine for enlarging the spirits when

you've just missed a rare DX station. Seriously though, a magnifying glass has a lot of uses, and I'll bet only a few amateurs even have one in the house.

With each issue of amateur radio magazines equipment seems to be smaller and smaller. Sometimes I think the people who put the equipment together must have magnifying eyeballs. Just look inside the gear and try to find the readings on a tiny transistor, a silicon diode, or even the colors on a small resistor without something to bring it up to a goodly size.

At one time, a magnifying glass was a handy device to start fires when the sun was bright. Now I use it to stop fires—to find out which resistor has started to burn up. There's so much packed into a modern piece of amateur gear, without a magnifying glass I don't believe anybody except a nearsighted old man could trace out the connections and components.

The other day I was having trouble with a rotary switch buried under a chassis. The naked eye couldn't tell what was wrong, but when I applied a glass to my eyes, I could see the corrosion on the ring and contact points. I also saw a cold solder joint at one connection.

Don't sell the magnifying glass short. Get one, but get a good one. By that I mean one made out of glass. There are cheap plastic varieties on the market. These aren't worth much, because they are easily scratched and can be marred by the slip of a soldering iron. I recommend one with a handle and a glass at least 2½ inches in diameter. I wouldn't be without one now.

Gay E. Milius, Jr., W4NJF

modifying the ART-13 for noiseless cw operation

The ART-13 is an excellent cw transmitter, but many operators don't use it because of the noise made by the relays pulling in when operated on cw.

To silence the relays for cw operation with the ART-13 requires temporary removal of the red indicator light, I-101. It isn't necessary to unsolder the wires of I-101; simply remove it from the mounting hole in the transmitter front panel so it can be pushed aside to remove wafer "emission" switch S-110. (This is labeled "off-voice-cw-mcw" on the front panel.) Do not disconnect the wires soldered to wafer switch S-110. Remove mounting screws and spacers between the two wafers. Replace the original spacers with spacers that are about half as long. Replace the wafers in their original position on the switch start.

Now, add more spacers the same length as the new shorter spacers you've already installed. This leaves the switch shaft long enough to accommodate an additional wafer. Select the contacts on the new wafer that will make when S-110 is in the cw position and will not make on any other position.

Ground one side of the new wafer contacts to the chassis, and connect the other side to the very tip end section of J-102 microphone jack. (When the emission switch is in the cw position the tip end section of J-102 is grounded, pulling in and holding all relays, which is the same thing that happens when the microphone push-to-talk button is energized.) Replace S-110 and I-101 on the front panel.

The transmitter is keyed when the emission switch is in the cw position, because vacuum relay K-102 grounds the cathode of the 837; thus the cathode lead to ground through the contacts of K-102 must be broken through the key jack. This may be accomplished by disconnecting the cathode lead from the ground side of R-131, a 350-ohm 10-watt resistor located on the bottom side of the transmitter in the center section directly behind the auto-tune motor. (One side of the resistor is connected to S-114, which is located on the bottom side of the transmitter directly behind the high-frequency knob" labeled "coarse A.") Make sure you disconnect the lead from the ground side of the resistor. This lead goes through vacuum switch K-102.

Now, isolate J-103 key jack above ground. This may be done by drilling the key jack hole in the chassis a little larger to accommodate insulating washers. Connect a lead from the ground side of R-131 to one side of J-103.

Connect another lead from the other side of J-103 to the original lead that was disconnected from the ground side of R-131. Insulate the connection.

Now the cathode of the 837 can be grounded through a key, when the emission switch is in the cw position, without the distracting noise of banging relays. **Note of caution:** Approximately 75 volts are on the key when keying the transmitter. I suggest you use a vacuum-tube keyer or an electronic keyer to eliminate the shock hazard.

To operate the ART-13 in the voice position with this modification, a shorted plug must be inserted in key jack J-103.

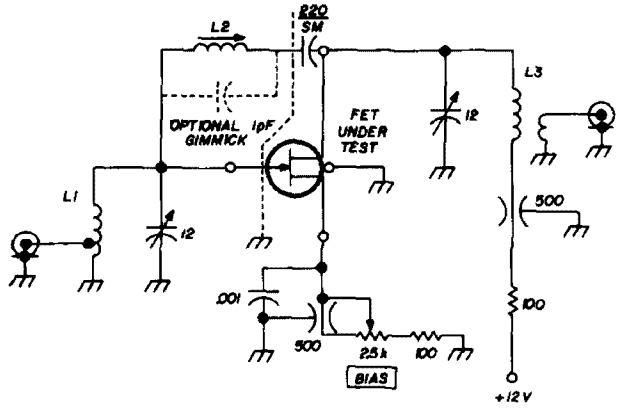
Don Whitney, K5GKN

choosing fets for 144 MHz converters

Here is some interesting information on jfets for two-meter converters. W6WSQ and I built a test jig (see **fig. 4**) for testing various jfets at 144 MHz. The test circuit is nothing more than a sensitive preamp which can be adjusted for different devices. A number of different fets were tried in the circuit; gain was measured, and weak signal performance was compared to a good low-noise vacuum tube preamp—the results are listed in **table 1.** The interesting thing is that sensitivity is just about the same for each device regardless of type. The only exceptions are the "first generation" fets (TIS34, 2N3823 and 2N2319), and their performance was only slightly lower than the best devices at 144 MHz.

Bruce Clark, K6JYO

fig. 4. Circuit of 144-MHz fet test jig. Reverse power supply voltage polarity for p-channel devices.



L1 5 turns no. 16, 3/8" diameter, 5/8" long, tapped at 1½ turns

L2 15 turns no. 26 enamelled on 3/8" slug-tuned form; brass slug. Form must be insulated from chassis.

L3 7 turns no. 18, 3/8" diameter, 3/4" long, secondary is 1½ turns no. 26 around cold end

table 1. Two-meter performance of various fet's.

type number	stable gain	weak signal source performance	manufacturer	price
2N5387	20-22 dB	note 1	Sil	$8.50
2N4416	15-18 dB	note 1	UC, TI	5.00
TIS88	15-18 dB	note 1	TI	1.75
A-2734	15-18 dB	note 2	Amelco	1.25
UC-734	14-18 dB	note 2	UC	1.10
IT-108	12-15 dB	note 2	Intersil	.85
TIS34	10-15 dB	note 3	Tf	1.10
2N3823	10-15 dB	note 3	TI	4.00
2N2319	10 dB	note 3	TI	—

note 1: 0.5 to 2 dB better than good 417A or 6CW4 units "normalized" at 3 dB noise figure

note 2: same or slightly better than comparison tube front ends

note 3: worse than comparison tube front ends; TIS34 to 1 to 2 dB worse, 2N3823 and 2N2319 3 to 4 dB worse

manufacturers: Sil, Siliconix; TI, Texas Instruments; UC, Union Carbide

mosfet transistors

In the cornucopia of mos field-effect transistors, I have selected a group which are most likely to be encountered by amateurs. A few types were omitted; perhaps a few were missed. In any event, the list should be an invaluable addition to your ham notebook. Because the emphasis in this grouping is on rf amplifier applications, all types are n-channel, depletion mode. This means their biasing and polarity are the same as vacuum tubes.

Don Nelson, WB2EGZ

table 2. Performance characteristics of n-channel mosfet's suitable for rf circuits.

type	gate	maximum freq (MHz)	application	trans-conductance (μmho)	200-MHz power gain (dB)	200-MHz noise figure (dB)	feedback capaci-tance (pF)	Price
3N128	single	250	rf amplifier, oscillator	7500	18	3.5	0.12	$1.45
3N139	single	250	video, af, rf amplifier with 35-V V_{dss}	6000	17	4.0	0.18	2.89
3N140	dual	300	rf amplifier with agc capability	8000	19	3.5	0.02	1.62
3N141	dual	300	converter, product detector	10,000	18[1]	—	0.02	1.55
3N142	single	175	rf amplifier, oscillator	7500	17[2]	2.5[2]	0.12	1.08
3N143	single	250	mixer, oscillator	7500	13.5[1]	—	0.12	1.39
3N152	single	250	premium rf amplifier	7500	20	2.5	0.12	1.78
3N159	dual	300	premium rf amplifier with agc capability	10,000	20	2.5	0.02	2.18
40467A	single	220	general purpose rf amplifier and oscillator	7500	16	4.5	0.12	1.24
40468	single	125	rf amplifier	7500	17[2]	3.5[2]	0.12	.75
40559	single	125	mixer, oscillator	2800[3]	21.5[1,2]	—	0.12	.70
40600	dual	250	*rf amplifier*	11,000	17.5	3.5	0.02	2.24
40601	dual	250	mixer	2700[3]	14[1]	—	0.02	1.16
40602	dual	—	i-f amplifier	10,000	28[4]	—	0.02	.83
40603	dual	150	rf amplifier	10,000	25[2]	2.5[2]	0.02	1.04
40604	dual	150	mixer	2800[3]	20[4]	—	0.02	.98
40673	gate-protected version of 3N140							2.22
MFE3006	dual	300	rf amplifier	15,000	24[2]	3.0[2]	0.02	1.50
MFE3007	dual	300	rf amplifier	14,000	24	3.0	0.02	1.65

1. Conversion gain
2. At 100 MHz
3. Conversion transconductance
4. At 44 MHz

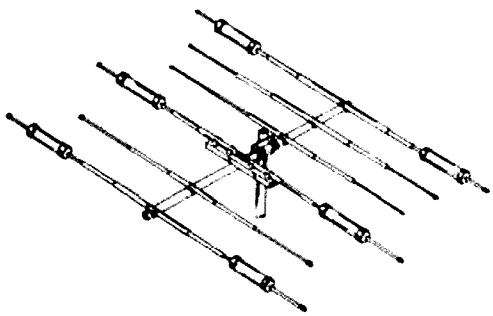
another use for coax relay coils

If you are using surplus coax relays with 24- or 28-volt dc coils, you might be able to use this idea. On a homebrew transmitter using push-pull 6AQ5's (the tubes could just as well have 6V6's), I found that the relay coil resistance was about the same resistance as the modulator cathode resistance should be. So I used the relay coil as the bias resistor. This worked okay, since the power supply B— was open on receive and grounded on transmit. Don't forget to bypass the coil with a 25-volt, 25 μF capacitor.

Bill Eslick, KØVQY

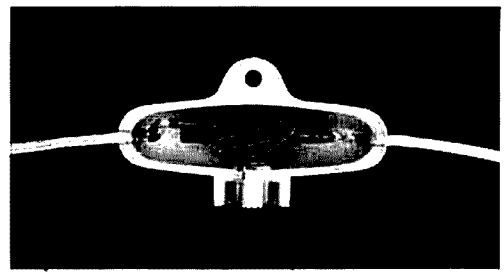
new products

mosley six-element tri-band beam



Mosley Electronics, Inc. has just announced their latest Trap-Master beam, the Classic 36, a six-element tri-band beam featuring the Classic feed system and balanced capacitive matching for efficient performance. The Classic 36 uses performance-proven Trap-Master traps for automatic bandswitching by means of high-impedance parallel-resonant tuned circuits. The new beam covers 10, 15 and 20 meters with an swr of 1.5:1 or better. Beam is power rated at 2 kW PEP ssb input to the final stage. If you already own a Mosley TA-36, you can obtain six-element tri-band operation with a TA36/CL36 conversion kit; kit includes an entire new radiating element. For more information write to Mosley Electronics, Inc., 4610 N. Lindbergh Boulevard, Bridgeton, Missouri 63042.

dipole center insulator



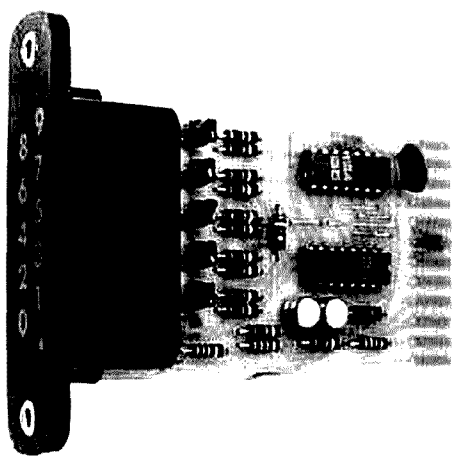
O. Watson Greene, W1CPI, has just introduced a new dipole center insulator that is available with or without a broadband balun. The housing for the Greene insulator is a precision molding of hard-flow polystyrene material. When the two halves are cemented together, they are practically impossible to separate. Rain and moisture are kept out of the insulator and feedlines by the rain drip collar that is formed around the type-UHF coaxial connector on the bottom of the insulator. The hoist ring at top center makes the unit adaptable for either inverted-vees or horizontal dipoles. The antenna connectors are quarter-inch copper braid, six inches long with tensile strength of 500 pounds. The 1:1 52-ohm balun is wound on a ferrite rod and covers the range from 2.8 to 32 MHz. The manufacturer reports that this balun will take the full legal amateur power input without core saturation. The model GWB with balun is priced at $10.00; the GNB without balun is $6.00. Order from O. Watson Greene, Wakefield, Rhode Island 02880.

noise blanker kit

The Drake 34-NB Noise Blanker Kit is designed for the Drake TR-3 and TR-4 transceivers. Unlike the noise clippers and limiters usually used in communications equipment, the 34-NB actually mutes the receiver for the duration of the noise pulse. Between noise pulses, full receiver gain is restored. The 34-NB is most effective on

strong, periodic noise impulses such as ignition noise and is least effective on random noise. However, impulse noise is the most troublesome, and loss of communications because of random noise is rare. According to the manufacturer, low-level signals that are completely masked by noise pulses without the noise blanker can be copied when it is used. This is a real boon to the mobile operator because he can blank ignition noise from trucks and other cars as well as his own car. $129 from your local distributor, or write to the R. L. Drake Company, 540 Richard Street, Miamisburg, Ohio 45342.

decade counter kit



Display Electronics has announced a new decade counter kit for the electronics experimenter. With these new digital modules, you can build professional looking frequency counters, digital meters, precision clocks and many other projects. The counter module uses Signetics "Utilogic" integrated circuits for high noise immunity and a guaranteed 10-MHz clock rate. Eleven neon lamps illuminate the ten digits and decimal point on the readout face.

The model DC-10-1 decade counter is available in kit form for $13.95 postpaid. The kit is easily assembled in a little over an hour and includes well written, easy-to-follow instructions. An assembled and tested version may be purchased for $16.95. Data sheets are available from Display Electronics, P. O. Box 1044, Littleton, Colorado 80120.

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FOR THE SWAN 250

The TNB-250 Noise Blanker effectively suppresses noise generated by auto ignitions, appliances, power lines, etc., permitting the recovery of weak DX and scatter signals normally lost in noise.

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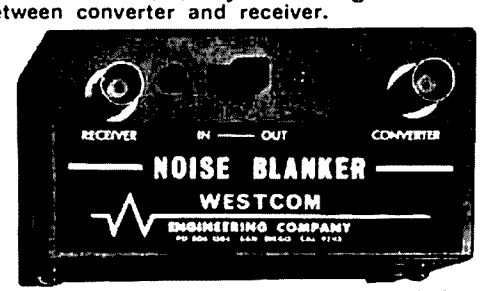


TNB-250 shown installed on a Swan 250 at accessory socket location.

Simplified installation requires twenty minutes.

TNB-250	$29.95 ppd.
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Refer to the New Products column of the August '68 issue of Ham Radio Magazine for additional information on the TNB Noise Blanker or write for technical brochure.

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low-cost volt-ohm-milliammeter

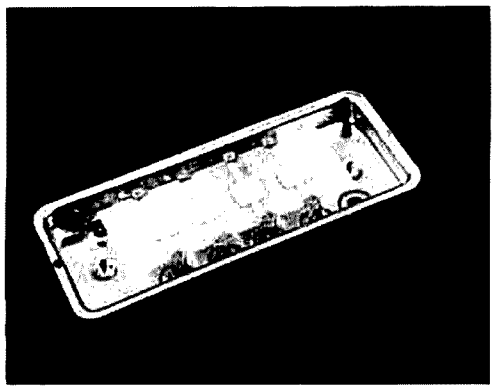


The first instrument new hams and electronic experimenters buy is a volt-ohm-milliammeter. It is the basic electronics test instrument and is an invaluable aid to any electronics trouble shooting. Among the functions of the vom are signal tracing and alignment, and voltage, current and resistance measurements. A new dependable pocket-size vom, the Knight-Kit model 646, is available in kit form at a low enough price for anyone interested in elec-

tronics. This instrument features an easy-to-read two-color scale, 20,000 ohm-per-volt dc sensitivity (10,000 ohms per volt on ac) and accuracy of ±2% full scale ac or dc.

The meter has five dc voltage ranges up to 1000 V, five ac ranges to 1000 V, three current ranges to 250 mA, three resistance ranges to 1 megohm, two dB scales plus scales for measuring large chokes and capacitors. Kit price is $11.95 with all parts, test leads and battery from Allied Radio Corporation, 100 N. Western Avenue, Chicago, Illinois 60680.

monolithic crystal filter



Tyco Laboratories has just announced a line of computer-designed high-frequency crystal filters with excellent performance characteristics. The center frequency range of these miniature filters is 5 to 22 MHz with fractional bandwidths from 0.01 to 0.4%. The new filters are currently available in a flat-pack metal can 1.54 inches long, 0.69 inch wide and 0.3 inch high. A typical eight-pole filter with a 10.7 MHz center frequency and 7 kHz bandwidth has a shape factor of 2:1 from 6 to 60 dB; stopband rejection is more than 100 dB, and spurious responses are at least 90 dB down.

Applications for these filters include ssb, CW and narrow-band fm communications systems. It's expected that the monolithic construction will offer significant cost reductions over discrete filters. For more information, write to Mr. Kenneth Thomson, Crystal Products Group, Tyco Laboratories, Inc., 1510 McGee Street, Kansas City, Missouri 64108.

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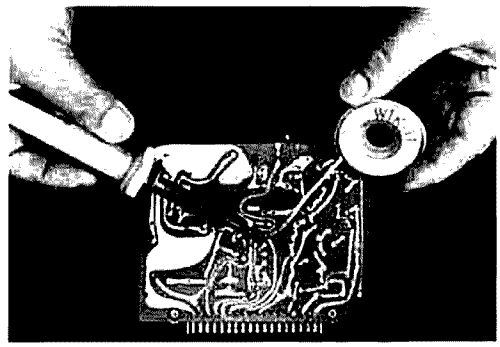
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desoldering tool



If you have ever tried to repair printed-circuit boards or tried to remove parts from them, here is a desoldering tool that should be of interest. Wik-It desoldering tool is a specially treated tinned braid that quickly draws up solder when heat is applied with an ordinary soldering iron. In addition to removing the solder, it acts as a heat sink to protect delicate components, and leaves the joint ready for re-soldering. It comes in various sizes and is sold in spools or as a service kit. See your local distributor or write to Wik-It Electronics Corporation, 2573 Spring Street, Redwood City, California 94063.

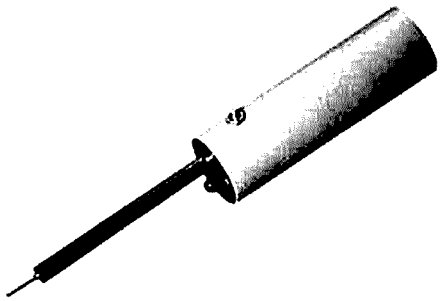
meter expander



A new meter expander that increases the voltage and current sensitivity of a vom by as much as a thousand times has just been announced by Integrated Controls, Inc. The meter expander has selectable gain settings of 10, 100 and 1000 and is connected as a buffer amplifier between the test leads and the vom. The direct-coupled IC buffer amplifier produces a nominal full output of ±2.5 volts and increases the voltage sensitivity of a vom up to 60 dB; input impedance is increased

to more than 1 megohm. It can also amplify current, multiplying the basic meter sensitivity up to 200 times. The gain-bandwidth product of the amplifier is 1 MHz. The unit is completely battery powered and its accuracy is compatible with most volt-ohm milliammeters. For more information, write to Integrated Controls, Inc., P. O. Box 17296, San Diego, California 92117.

signal-injector kit



If you've ever done any troubleshooting, you know one of the handiest gadgets you can use is a signal injector. The new Knight-Kit KG-644 solid-state signal-injector kit is completely portable and self-contained and produces a signal rich in harmonics for tracing audio, rf and i-f circuits. The test probe is insulated for maximum safety, and the unit has a built-in battery-condition light. The kit is priced at $4.95, complete with 4 penlight batteries, from Allied Radio Corporation, 100 N. Western, Chicago, Illinois 60680.

varitronics correction

Unfortunately, two typographical errors crept into the new products writeup on the new FDFM-2S fm transceiver manufactured by the Inoue Communications Equipment Corporation. The power supply can vary between 12 and 14.5 Vdc, and frequency deviation is less than 15 kHz. Since the new products release that appeared in the June, 1969 issue of **ham radio**, the price of this transceiver has been increased to $310. The unit is distributed by Varitronics, Inc., 3835 North 32nd Street, Suite 6, Phoenix, Arizona 85018.

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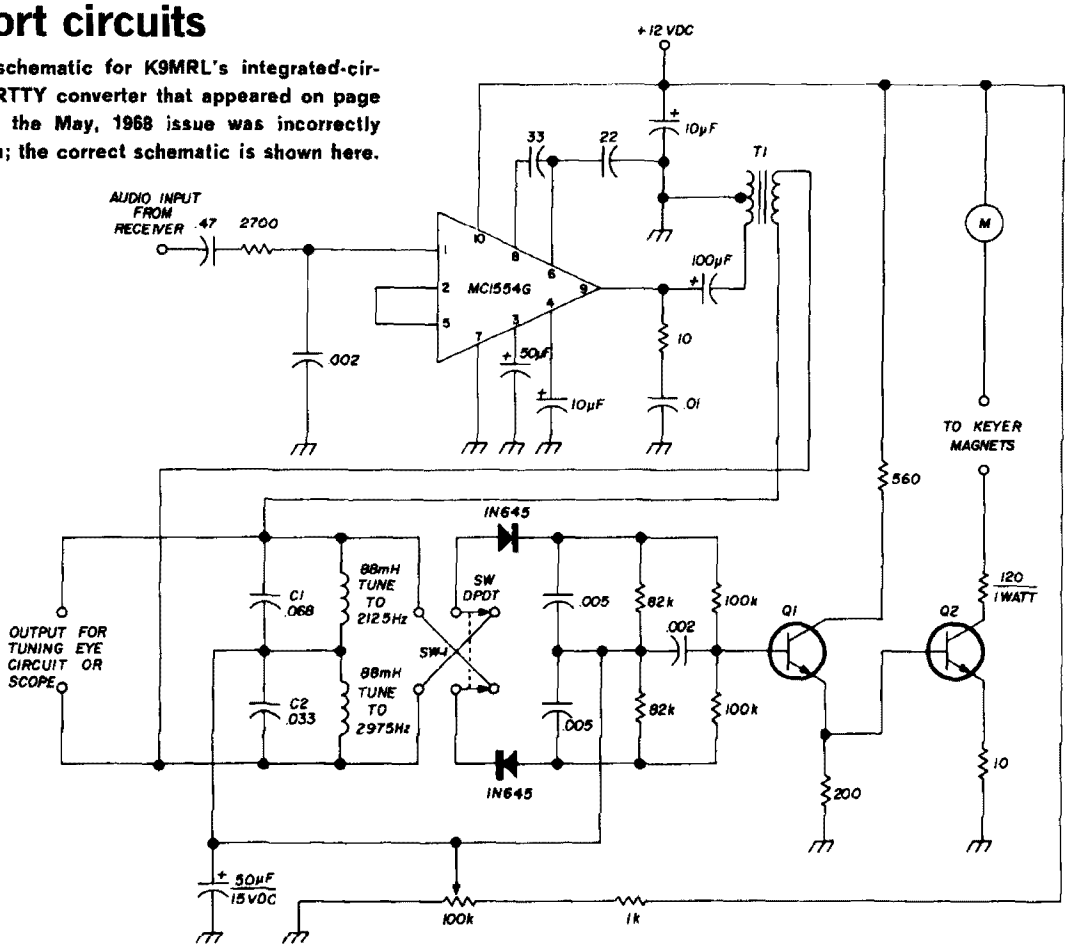
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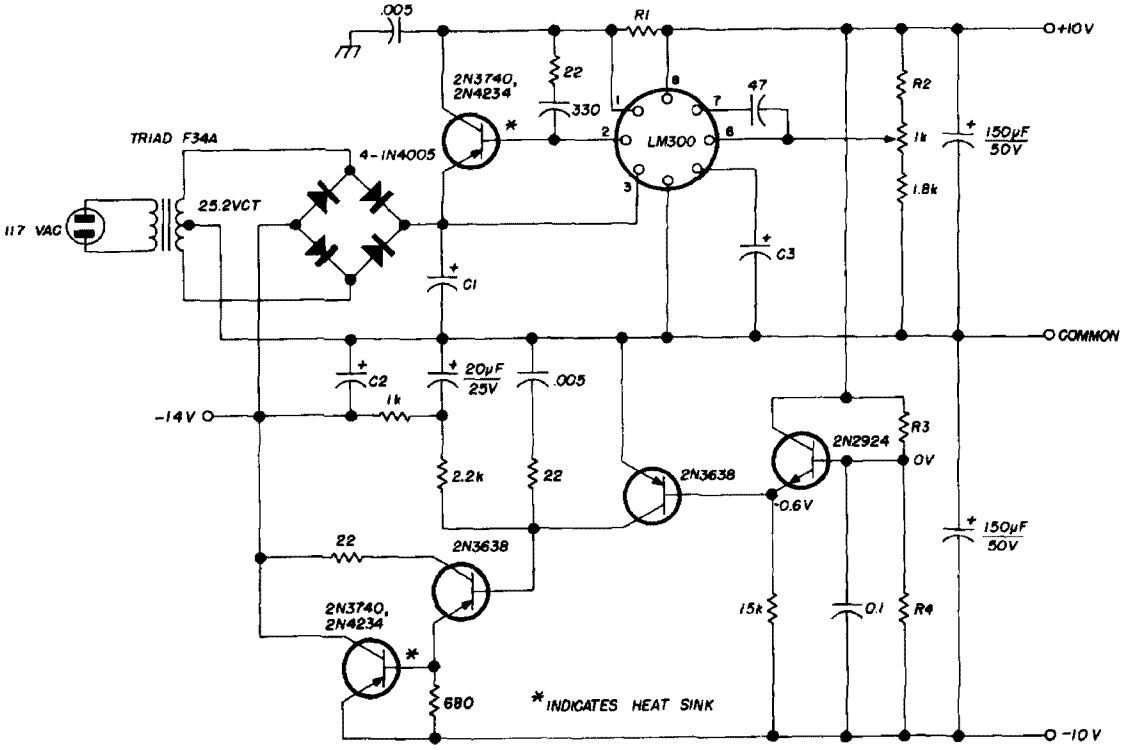
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short circuits

The schematic for K9MRL's integrated-circuit RTTY converter that appeared on page 41 of the May, 1968 issue was incorrectly drawn; the correct schematic is shown here.



The schematic for W1OOP's slaved dual-voltage power supply that appeared in the ham notebook section of the April issue was in error—the collector circuit of the 2N2924 should be connected to the positive supply. Resistors R3 and R4 should be 1 percent resistors; any value fom 2k to 22k.

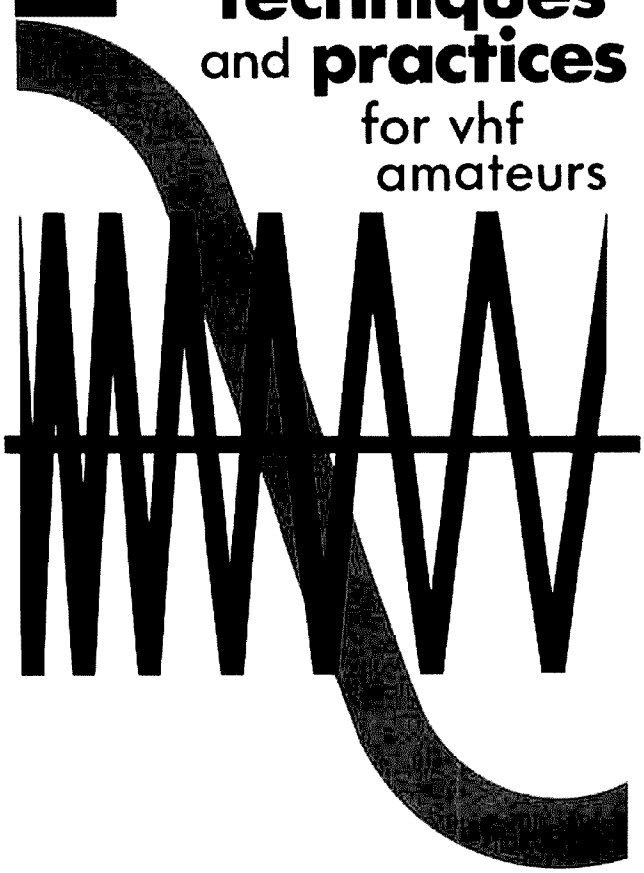


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fm techniques and practices for vhf amateurs



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september 1969

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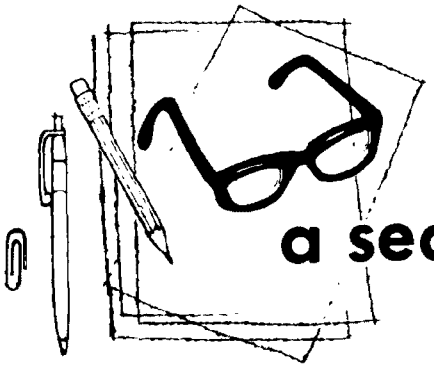
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a second look

by jim fisk

One of the things that makes amateur radio stand out among other recreational pursuits is the diversification it offers. Many hams like to relax after a hard day's work and talk to friends, handle traffic, or chase DX. To others, amateur radio is a means for intellectual development and personal achievement. Some fellows have a drive that makes them explore beyond the canned circuit and ready-made construction project. They like to improve, modify, innovate. These are the experimenters, and their greatest satisfaction comes from finding new ways to do things. They must know what makes things tick and why. The experimenters are truly responsible for the perpetuation of our hobby. Without them, it's doubtful if amateur radio would have existed as we know it today.

If you're a serious experimenter, ham radio can be a means of enhancing your advancement in electronics. An example is the ham who is also an engineer. When solving a design problem at work, he may remember something he's read that somehow correlates with his immediate task. He may even work on the idea at home. Often these experimenters make significant contributions to the advancement of electronics—you see their results published in the literature.

One of the many areas open to experimenters is medical electronics. Doctors working on research projects in the medical disciplines can use the help of electronic engineers and technicians. For example, a physician working in radio isotope research needed an interface between his computer and a photoelectric sensor. This would be a fairly simple task for one familiar with digital circuits. The doctor needed a buffer memory and an analog-digital converter. He couldn't find anyone to help him, so he boned up on theory and actually built the thing himself—it worked!

Remember the tunnel diode? Engineers in the Device Research Branch of NASA's Electronic Research Center found that pressure applied to a tunnel diode made of gallium antimonide caused the diode's band gap to vary. Apparently, pressure differential in this region changes the inflection point on the diode's characteristic current-voltage curve. The resulting change in negative resistance (and hence maximum gain) makes the device useful as a transducer.

The NASA research people built a tunnel diode transducer using these principles. The transducer was about 0.05 mm thick—small enough to pass through the eye of a needle. Power requirements were 50 μW with 50 mV input. When inserted into the cardiovascular system, the diode reproduces minute changes in blood pressure. To the physician, these variations reveal essential data on heart activity and the circulatory system. The transducer's extremely small size allows doctors to monitor blood-pressure parameters heretofore undetectable with conventional sensors.

These are but a few things open to electronic experimentation. You probably won't come up with anything as exotic as a cardiovascular transducer, of course, but surely you must have some ideas that are worth developing in your home workshop. For example, how many different applications can you think of for the IC operational amplifier? Think about it, work up a design, then heat up your soldering iron.

Jim Fisk, W1DTY
editor

fm techniques and practices for vhf amateurs

William I. Orr, W6SAI, Eimac Division of Varian, San Carlos, California 94070

In 1937 the residents along the New York and New Jersey sides of the Hudson River were awed and mystified by the appearance of a gigantic spidery structure that arose near Alpine, New Jersey. It was whispered by the local wags that this tower transmitted "secret waves," and rumors flew that the tower was used to contact inhabitants of Mars, that it was used for death-ray experiments, and that it was used for secret communications with the hereafter.

None of these wild stories, of course, were true. The appearance of the tower was the first public knowledge of a new mode of radio communication, based upon experiments of Major Edwin H. Armstrong of Columbia University, an early wireless pioneer and inventor of the regenerative receiving circuit.

static-free radio

Major Armstrong had long realized that amplitude-modulated broadcasting suffered badly from static interference generated by electrical machinery and from crashes and noises created by natural electrical discharges in the atmosphere. It was Armstrong's belief, supported by the 1922 mathematical studies of John Carson,[1] that the way to avoid static was to artificially produce a radio signal differing as radically as possible from the static signals that wreaked havoc with a-m broadcast transmissions. If this could be done, a receiver could be built which would be insensitive to static interference.

Armstrong reasoned that random noise and static were amplitude variations that had no orderly variations in frequency. He proposed that a receiver be built that would be sensitive to frequency shifts but insensitive to amplitude variations. He started to work on such a receiver, plus a companion transmitter to generate a **frequency modulated** (fm) signal. During the course of his experiments in the Alpine, New Jersey laboratory, a 400-foot tower was erected. This impressive structure, with three large crossarms, was visible for miles along the banks of the Hudson River.

In 1934, before the tower was built, Armstrong conducted preliminary experimental frequency-modulated tests with a 50 watt fm transmitter atop the Empire State Building. These tests were run in conjunction with W2AG, Yonkers, New York, who operated a 50-watt fm transmitter at 120 MHz in the old $2^1/_2$-meter amateur band. Voice, music, facsimile and multiplex telegraphy were transmitted by fm to W2AG.

In 1935, encouraged by the results of the tests, Armstrong applied to the FCC for an experimental fm license and was granted the call W2XMN for use in the experimental frequency region around 40 MHz. Armstrong's fm tests were soon followed by others, particularly by Connecticut State College, operating stations W1XCS and W1XPW near 43 MHz. Finally the FCC set aside four channels—200 kHz bandwidth each—for exclusive fm broadcasting in the 40-MHz band. Commercial fm broadcasting was finally launched, although it was nearly eclipsed by World War II and the advent of commercial television.

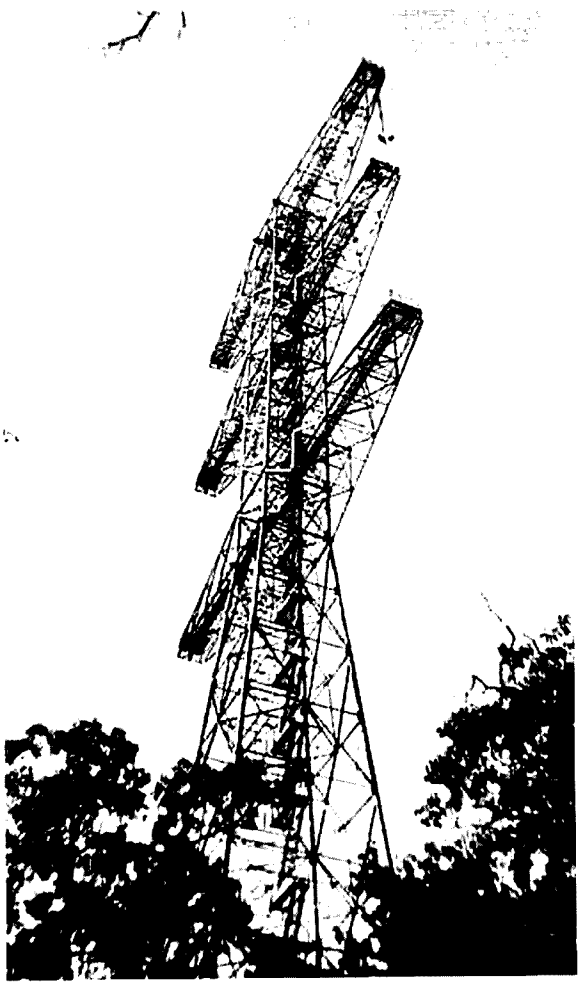
Radio amateur interest in fm techniques was aroused by the W2AG tests, and during the period from 1935 to 1940 a series of classic articles in *QST, Radio* and *Electronics*[3-8] provided practical circuitry for the

transmission and reception of fm signals in the amateur 5-meter band. Amateur experiments before the war showed that fm promised excellent prospects for static-free, reliable mobile communication on the "ultra high" frequencies of 2½ and 1¼ meters.

post war fm

Frequency modulation proved its worth during World War II when it was extensively used for vhf vehicular communication and over-the-horizon military relay circuits. Unfortunately, the times were against amateur fm usage in the late 1940's. Aside from sporadic use on the higher frequencies, fm languished in the amateur bands. In the vhf bands where it would have been effective, fm was obliterated by the flood of surplus a-m gear that invaded the market.

The tower at Alpine, New Jersey as it appeared in 1940. The man working on the 42.8-MHz turnstile antenna is supposedly Armstrong in his bucket sling.



The SCR-522 a-m transmitter-receiver was king as far as the new 2-meter band was concerned. Aside from a few eccentrics, vhf operators restricted themselves to a-m, with a sprinkling of cw for DX work. Some narrow-band fm experiments took place on the hf amateur bands, but nbfm never proved popular except as a means for escaping broadcast interference. By 1950 or so, amateur fm was largely forgotten.

While amateurs limited their efforts to a-m in the world above 50 MHz, the mobile radio service, quick to see the advantages of fm, made a complete conversion to that mode, using 50-kHz channels and crystal-controlled transmitters and receivers. Eventually, the scarcity of channels in the mobile radio service spectrum became so acute that channel spacing was halved to 25 kHz. This released a flood of obsolete mobile fm gear to the amateur surplus market.

Amateurs soon realized the inherent advantages of this "new" form of vhf communication, and today the use of fm is increasingly popular on the higher frequency bands. Indeed, I predict that fm will eventually supplant a-m and ssb on the 144-, 220- and 432-MHz amateur bands in the metropolitan areas of the United States and Canada. The application of fm techniques in combination with stationary, remote repeater systems will bring about a revolution in the operating habits of vhf oriented amateurs during the coming decade.

This article covers frequency modulation fundamentals and techniques and discusses the use of fm equipment and repeaters. It is hoped that it will serve as an introduction and stimulant to amateurs who have not yet experienced the thrill of using vhf fm communication as a means of reliable short range communication.

fm: what is it?

Intelligence may be superimposed upon a steady radio signal (carrier wave) by altering the structure of the wave in some way. The amplitude of the wave may be changed as is done in amplitude modulation or ssb. Or, the frequency, or phase, of the wave may be varied. In either case, the process of altering the radio wave in accord with the transmitted

intelligence is called **modulation.** Modulation imparts a distinguishing characteristic to the radio wave which enables a properly designed receiver to convert it back into a replica of the original intelligence.

Fig. 1A shows an rf carrier that is amplitude modulated by a sinusoidal audio signal. The modulated rf wave varies about the zero axis at a constant rate, and the strength of the individual cycles of the wave is proportional to the amplitude of the modulating signal. The carrier is modulated 100 percent when the peak-to-peak amplitude of the modulating signal is equal to the peak-to-peak amplitude of the unmodulated carrier.

In **fig. 1B** the rf carrier is frequency and phase modulated by the same audio signal. It can be seen that modulation voltage of one polarity causes the carrier frequency to decrease, as shown by the fact that the individual rf cycles are spaced farther apart. A modulating voltage of the opposite polarity causes the frequency to increase; this is shown by the fact that more rf cycles are completed in a given time interval. If the phase-modulated (pm) wave is shifted 90°, the two waves in **fig. 1B** look alike. No amplitude variation takes place in either wave.

When an rf carrier is amplitude modulated, additional side frequencies are generated. These frequencies (sidebands) are located on each side of the carrier, spaced from the carrier by an interval equal to the modulation frequency. The strength of the carrier does not vary during modulation, but the strength of the sidebands depends upon the percentage of modulation. When the carrier is modulated 100 percent, the power in the sidebands is equal to one-half that of the carrier.

When an rf carrier is frequency modulated, many more than two additional side frequencies are generated. The first two are spaced from the carrier by the modulation frequency; additional side frequencies are located on each side of the carrier, spaced from each other by an amount equal to the modulating frequency. Theoretically, there are an infinite number of side frequencies; but, practically, the strength of those beyond the frequency swing of the modulated carrier is relatively low.

Unlike amplitude modulation, the strength of the carrier component of an fm wave varies widely; under certain conditions it may even disappear entirely. This variation in the strength of the carrier component is useful in measuring the modulation index, as will be discussed later.

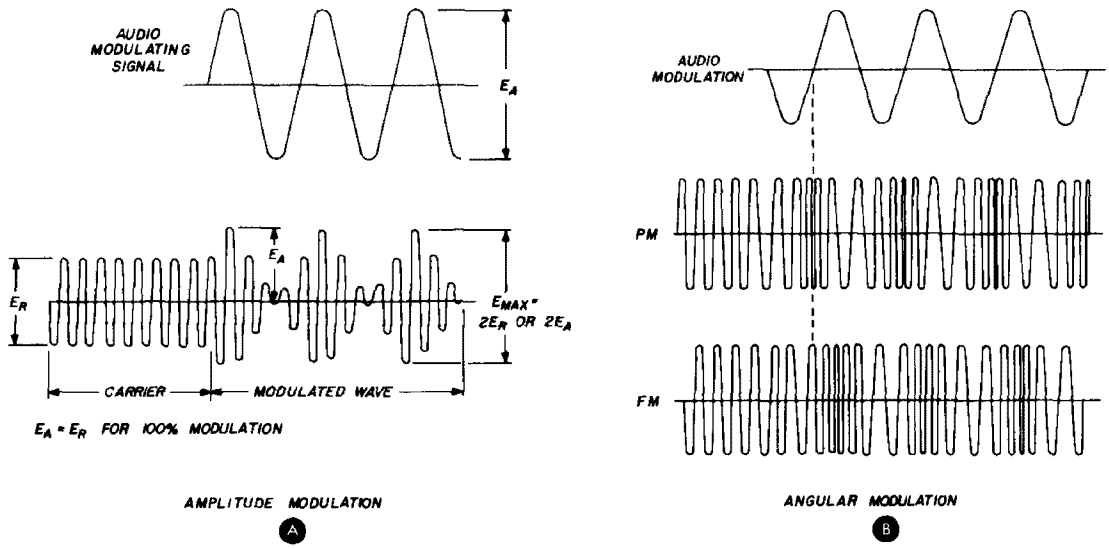


fig. 1. Amplitude modulation (A) and angular modulation (B) of an rf carrier by the same audio signal.

fm terms

There are several terms which convey considerable information about the character of

a frequency-modulated signal: deviation, swing, modulation index and deviation ratio.

Deviation specifies the amount of frequency shift to each side of the unmodulated carrier when the transmitter is modulated. Deviation is directly proportional to the amplitude of the modulating signal and is usually measured in kilohertz. The limits of frequency shift on either side of the carrier are known as the frequency deviation limits.

Swing is the total frequency range covered by an fm transmitter when modulated with a symmetrical signal (equal deviation to each side of the carrier frequency). If, for example, a transmitter operating on 50,000 kHz has its frequency shifted to 50,010 kHz, back to 50,000 kHz, then to 49,990 kHz, and back again to 50,000 kHz during one cycle of the modulating wave, the deviation is 10 kHz and the swing is 20 kHz.

Modulation index of an fm signal is the ratio of the maximum frequency deviation to the audio modulating frequency, when both are expressed in the same units. Thus, in the previous example, if the signal is varied from 50,000 kHz to 50,010 kHz to 49,990 kHz and back to 50,000 kHz at a rate (frequency) of 2000 times per second (modulation frequency = 2 kHz), the modulation index would be 5, since the deviation (10 kHz) is 5 times the modulating frequency (2 kHz). Expressed as a formula, the modulation index is:

$$\text{Modulation Index} = \frac{\text{Maximum Frequency Deviation}}{\text{Maximum Frequency of modulating signal}}$$

Deviation ratio is similar to the modulation index in that it involves the ratio between the modulating frequency and deviation. In this case, however, the deviation in question is the peak frequency shift obtained under full modulation, and the audio frequency considered is the highest one transmitted by the system. For example, if the maximum audio frequency to be transmitted is 3000 Hz, a deviation ratio of 3 would call for a peak deviation of 3 x 3000, or 9 kHz at full modulation.

The noise suppression capabilities of fm are directly related to the deviation ratio. When the deviation ratio is increased, noise suppression is better when the signal is stronger than the noise. When the noise approaches the signal in strength, however, lower deviation ratios allow communications to be maintained in many cases where high-deviation-ratio fm is incapable of giving service. This assumes that a narrow-band fm receiver is used.

For each value of received signal-to-noise ratio there is a maximum deviation ratio. Beyond this maximum deviation ratio the output audio signal-to-noise ratio decreases. Up to this critical deviation ratio, noise suppression becomes progressively better as the deviation ratio is increased.

For high-fidelity fm broadcasting, a deviation ratio of 5 is ordinarily used. The maximum audio frequency is 15 kHz and the peak deviation at full modulation is 75 kHz. For tv sound, the deviation ratio is 1.67, the maximum audio frequency is 15 kHz, and the peak deviation is 25 kHz.

Narrow-band fm has been standardized by the public services and business radio service. A maximum deviation of 5 kHz is used by these services in the vhf region with 25 kHz channel separation. In the uhf region a deviation of 15 kHz is used along with 50 kHz channel separation.

In the amateur service, the bandwidth of an fm signal on frequencies below 29 MHz and between 50.1 and 52.5 MHz cannot exceed that of an a-m signal having the same audio characteristics [Section 97.65(c)]. Thus, wideband fm can be used above 29 MHz on 10 meters, above 52.5 MHz on the 6-meter band and in all the amateur vhf and uhf bands. Most fm stations in the vhf amateur bands use deviations between 5 and 25 kHz, with a trend toward the lower values.

fm sidebands

I mentioned earlier that additional side frequencies over and above those generated by a-m are created by frequency modulation. For distortionless transmission and bandwidth conversion a study of the spectrum of a frequency-modulated wave is useful. Mathematical studies and observation of the fm sidebands with a wave analyzer can provide

information that will allow efficient modulation under conditions of maximum deviation and minimum bandwidth.

The mathematical expressions for an fm or pm signal may be written as an infinite series having coefficients known as Bessel functions, named after a German mathematician who first studied variations of this kind. A physical representation of an fm or pm signal may be achieved by the use of a vector. The vector may be drawn as an arrow, with its length denoting the magnitude of the voltage and the arrow indicating direction. The position of the vector with respect to various reference axes can be used to indicate angular displacement as is commonly done in vector representations of electric and magnetic fields. The frequency of an alternating voltage represented by a vector, for example, is proportional to the speed of rotation of the vector with respect to time.

The frequency-modulated carrier shown in **fig. 1B** may also be represented by visualizing the carrier as a vector rotating counterclockwise with time (**fig. 2A**). For each cycle of the sinusoidal carrier, the vector rotates one complete circle or 360°. This is the period of the wave; the vector rotates at a frequency equal to the wave it represents.

For the case of amplitude modulation, the vector rotates at the carrier frequency but alternately changes amplitude (at the modulation frequency) about its unmodulated value. Since the carrier frequency is much higher than the modulation frequency, the vector undergoes many rotations during each modulation cycle. As a result, the tip of the vector moves alternately inwards and outwards over a close multiturn spiral path.

Since the angular position of the vector is unaffected by amplitude modulation, the observer may ignore its rotation by considering it from the point of view of an observer who is rotating with it, and restrict his attention to its changing length. For a-m then, the representation can be reduced to a line which expands and contracts at the modulation frequency and reverses its direction each time the modulation goes through zero.

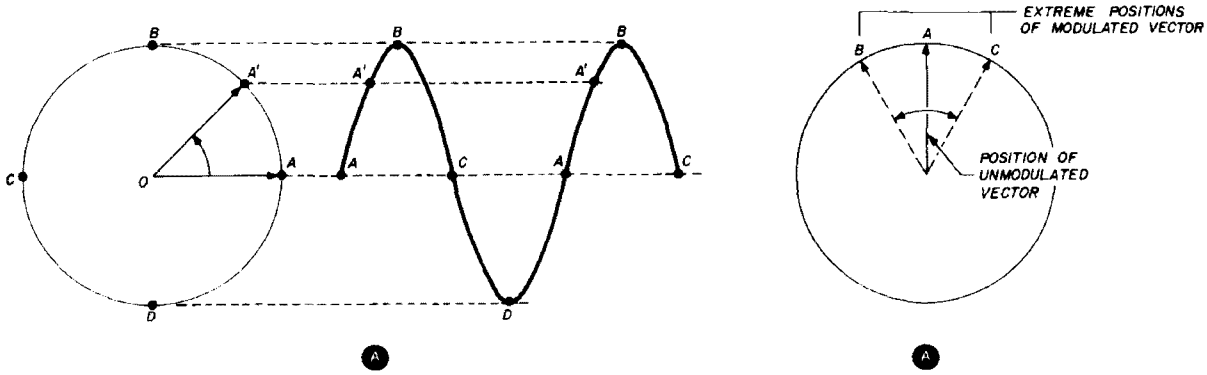


fig. 2. A rotating vector may be used to represent a sinusoidal wave (A). An fm signal may be regarded as a constant-length vector rotating with slowly varying speed; the angular displacement in advance or retard of normal is caused by angular modulation (B).

When the amplitude of the carrier vector is held constant, modulation may only be accomplished by changing the otherwise uniform rate of rotation of the vector. A constant amplitude, variable frequency signal can be represented by a vector of constant length, rotating with slowly varying speed. When the angular velocity of a vector is varied higher or lower about its mean value, its angular position is advanced or retarded from the angular position the unmodulated vector would have occupied (**fig. 2B**). Therefore, variations in angular velocity produce an alternate advance and retardation in the phase of the carrier.

Conversely, if it were desired to advance the angular position of the rotating vector by a certain amount it could only be done by temporarily increasing its angular velocity (frequency). Likewise, a backward shift of

the rotating vector could only be produced by reducing the frequency from the normal value. To preserve any lead or lag that was introduced, the frequency would be restored to its normal value.

Modulation based upon the advancement or retardation of the angular position of a rotating vector representing the carrier wave is called **angular modulation.** Two practical forms of angular modulation are **frequency modulation** and **phase modulation.**

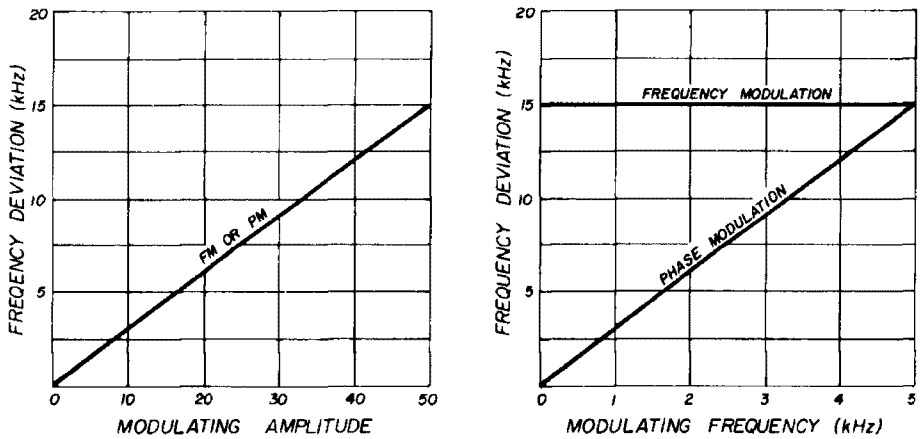


fig. 3. Relationships between fm and pm. The only difference between pure fm and pm is the fact that the deviation is a function of the amplitude only of the modulating signal for fm while the frequency of the modulating signal also determines the deviation for pm.

frequency and phase modulation

When frequency modulation is used, the phase of the carrier wave is indirectly affected. Similarly, when phase modulation is used, the carrier frequency is affected. You cannot vary frequency without changing phase and vice-versa. In view of the interrelationship between frequency and phase, with the two quantities changing together under angular modulation, the distinction between frequency and phase modulation is only a nominal one.

The only difference between pure frequency modulation and phase modulation is the fact that deviation is a function of **amplitude only** of the modulating signal for pure frequency modulation, while the **frequency** of the modulating signal **also** determines the deviation for phase modulation. A comparison of the two signal characteristics is shown in **fig. 3.**

Most modern fm transmitters use phase modulation, because frequency modulation can only be applied to an oscillator stage, while phase modulation may be readily applied to any amplifier stage. Therefore, phase modulation is easily applied to crystal-controlled transmitters. With phase modulation, if an audio signal of 1000 Hz causes a deviation of 0.5 kHz, for example, a 2000 Hz modulating signal of the same amplitude will produce a deviation of 1 kHz, and so on. To produce an fm signal with this technique, it is only necessary to make the deviation independent of the frequency of modulation, and proportional only to the amplitude of the modulating signal. This is accomplished by including a frequency correcting network in the transmitter.

The only disadvantage of pm, as compared to direct fm, is the fact that very little frequency deviation is produced directly by the phase modulator.* The degree of deviation is dependent only upon the phase deviation produced on the modulation frequency:

$$F_d = M_p$$

*This is the only obvious practical disadvantage. In reality, phase modulation suffers one other serious disadvantage. In pure fm, the spectrum width can be made nearly independent of the modulating frequency; this is not true with pure pm. The undesirable variation of bandwidth with modulating frequency in phase modulation can be corrected with a tailored audio clipper. The clipper is both pre-and post-equalized: audio with a 6-dB-per-octave rising characteristic is fed to the clipper; clipper output is then passed through a network with a 6-dB-per-octave falling characteristic. The processed audio going to the phase modulator has a flat response, but with tailored clipping which results in constant peak deviation after modulation.

where F_d is the frequency deviation one way from the mean value of the carrier, and M_p is the phase deviation accompanying modulation (expressed in radians per second). Note that the deviation is completely independent of the carrier frequency. The amount of phase shift that can be obtained with good linearity is such that the maximum practical modulation index is about 0.5.

deviation of 625 Hz at a crystal frequency of 6 MHz: this is well within the linear capability of a phase modulator. Some high-frequency fm gear designed for operation from 30 to 50 MHz uses crystals in the 200- to 500-kHz region to allow for sufficient frequency multiplication for satisfactory wide-band phase modulation at the carrier frequency.

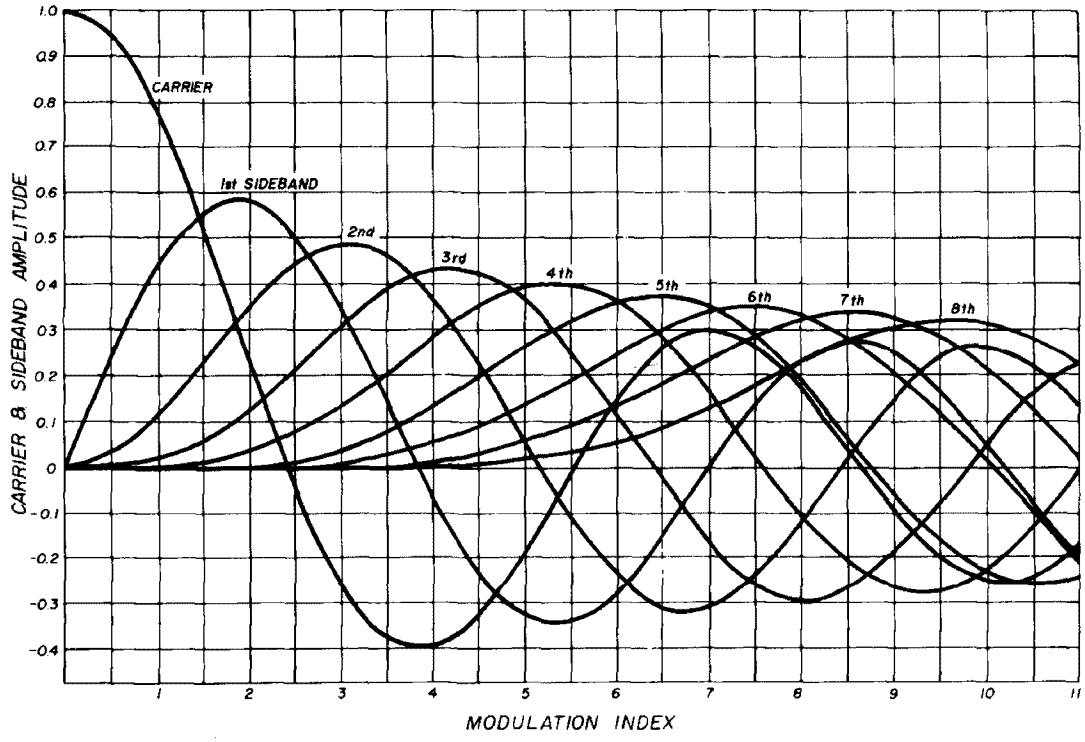


fig. 4. Bessel curves show variation in carrier and sideband amplitude with modulation index.

Zero Carrier Point	Modulaton Index (m)	To Obtain Bandwidth Multiply Modulating Frequency By
1	2.40	6
2	5.52	8
3	8.65	12
4	11.79	14
5	14.93	16
6	18.07	18
7	21.21	20
8	24.35	22
9	27.49	26
10	30.64	28

The degree of frequency deviation achieved by phase modulation can be increased by multiplying the signal frequency **after** modulation. Since there are no amplitude variations in any form of angular modulation, the signal may be amplified by a nonlinear stage such as a frequency multiplier or a class-C amplifier. Modulation can take place at a low level and at a low frequency, and then be amplified by frequency multipliers or straight amplifiers. The modulation index of the signal is multiplied by the same factor by which the carrier frequency is multiplied.

Many vhf fm transmitters are crystal controlled by a crystal that is 1/24 or 1/32 of the carrier frequency. A deviation of 15 kHz at 144 MHz, for example, is equivalent to a

Odd-harmonic distortion is produced when fm is obtained by the phase modulation technique, and the amount of distortion that can be tolerated determines the amount of phase modulation that may be used. Since the audio-frequency correction network causes the lowest modulating frequency to have the greatest amplitude, maximum phase

modulation takes place at the lowest modulating frequency, and the amount of distortion that can be tolerated at this frequency determines the maximum deviation that can be obtained by the phase-modulation technique. Normally, for vhf fm, the deviation ranges from 5 kHz to 25 kHz. With a maximum audio frequency of 3000 Hz for voice operation, this represents a maximum deviation of about 8. Therefore, low-frequency crystals followed by multiplier stages are mandatory for this type of service.

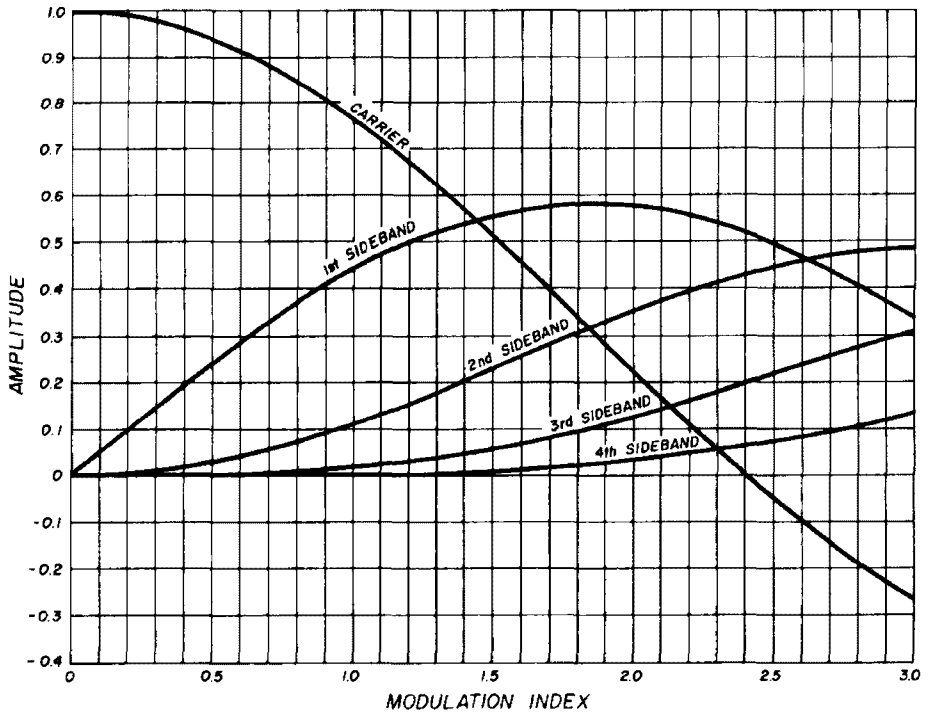


fig. 5. An expansion of the Bessel curves of fig. 4 for modulation indexes up to three.

fm sidebands

Until mathematical analysis proved otherwise, some people believed that varying the carrier frequency by an amount less than the actual audio frequency (for example, causing a 10-kHz modulating signal to vary the carrier frequency by plus or minus 1 kHz) would allow transmission over a frequency band smaller than that required by amplitude modulation. However, no matter how small the frequency swing in fm may be, the upper and lower sideband frequencies that are produced in a-m are also present in fm.

Moreover, not only are these two side frequencies present, but also present are an infinite number of pairs of such frequencies mathematically related to the frequency of modulation. Fortunately, only a limited number of these pairs (those nearest to the carrier frequency) are of significant amplitude for concern. However, when Armstrong first conducted his experiments with wideband fm, he used a frequency spectrum of about 250 kHz for high-fidelity transmission of music. For this reason modern fm broadcasting must be carried out in the vhf region, since it is only here that sufficiently wide channels are available.

The sidebands generated in fm differ from those resulting from a-m in that they occur at integral multiples of the modulating frequency; in a-m a single set of sidebands is generated for each modulating frequency. A simple method of determining the amplitude of the various f-m sidebands is the family of Bessel curves shown in **fig. 4.** There is one curve for the carrier and one for each pair of sideband frequencies.

The Bessel curves show how the carrier and sideband frequency pairs rise and fall with increasing modulation index, and illustrate the particular values at which they disappear as they pass through zero. Since the

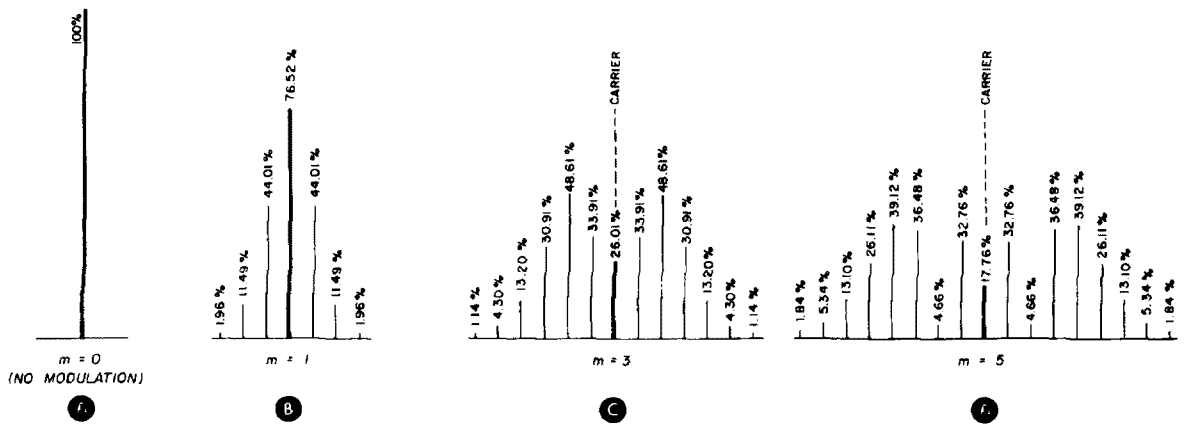


fig. 6. Effect of modulation index showing sideband frequency amplitude and distribution for modulation indexes 1, 3 and 5.

carrier drops to zero at particular values of modulation index (2.4, 5.5, 8.7, 11.8, etc.), the modulation index can be measured by noting the disappearance of the carrier. An expanded set of Bessel curves for modulation indexes up to 3 is plotted in **fig. 5.**

The relative amplitudes of carrier and sideband frequencies for any modulation index can be determined by finding the y-axis intercept for the particular function. Representative spectrum plots for three different values of modulation index are shown in **fig. 6.** The negative amplitude in the Bessel curves indicates that the phase of the particular function is reversed as compared to the phase without modulation. In fm, the energy that goes into the sideband frequencies is taken from the carrier; the total power in the over-all composite signal remains the same regardless of the modulation index.

It should be noted that the frequency spectrum for fm is relatively constant for a given modulation depth regardless of the modulation frequency. On the other hand, the bandwidth of a pure pm signal increases with both modulation amplitude and frequency.

This situation does not occur in amplitude modulation. You might think that the large number of sideband frequencies would make the frequency spectrum produced by an fm transmitter prohibitively

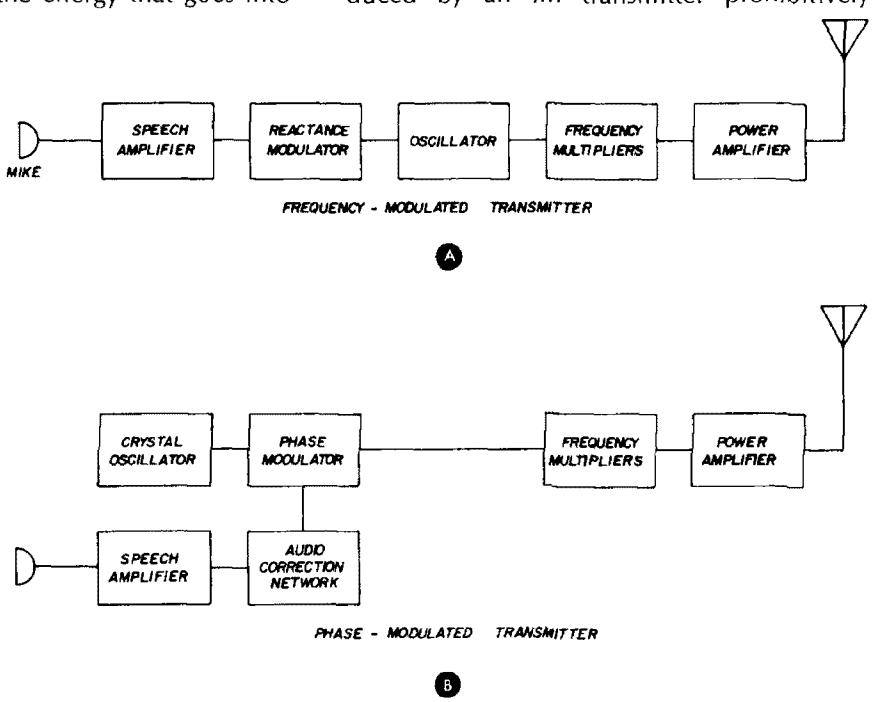


fig. 7. Block diagrams of fm and pm transmitters.

wide. However, the additional side frequencies are of small amplitude, and instead of increasing the bandwidth, modulation by a complex wave actually reduces the effective bandwidth of the fm spectrum. This is especially true when speech modulation is used, since most of the power in voice sounds is concentrated in the lower audio frequencies.

generation of fm signals

Frequency modulation can be obtained directly by changing the frequency of an

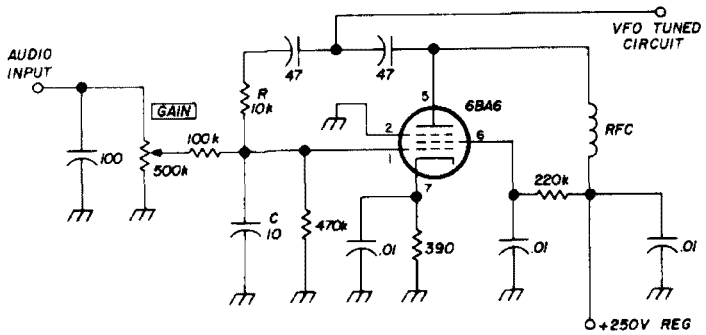


fig. 8. Simple reactance-tube modulator.

oscillator with the modulating signal, or indirectly with a phase modulator. Phase modulation circuits will be discussed later.

Simple frequency modulation of an oscillator avoids the need for phase- to frequency-modulation conversion. In addition, it produces a large proportional frequency change, so less frequency multiplication is needed.

A block diagram of an fm transmitter is shown in **fig. 7A**. In the fm transmitter, the output of the speech amplifier is usually connected directly to a modulator stage, which produces an equivalent reactance that varies with the modulating signal. This causes the frequency of the oscillator to vary in the same way. The frequency-modulated signal of the oscillator is then multiplied to the desired output frequency by the multipliers; a power amplifier boosts the signal to the desired output level.

the reactance modulator

The reactance modulator is a device that is connected across the oscillator tuned circuit to vary its resonant frequency in accord with applied modulation. A vacuum tube or transistor can be made to appear as a capacitive or inductive reactance by exciting the modulator with a voltage which either leads or lags the oscillator tank voltage. This leading or lagging input voltage causes a corresponding leading or lagging current, and the output circuit appears as a capacitive or inductive reactance across the oscillator tank circuit.

The transconductance of a vacuum-tube reactance modulator has an inverse effect on the impedance across the oscillator tank circuit. Since transconductance changes with grid voltage, the magnitude of the reactance across the tank circuit is varied with the grid signal of the reactance tube. If the impedance is complex (composed of reactance and resistance)

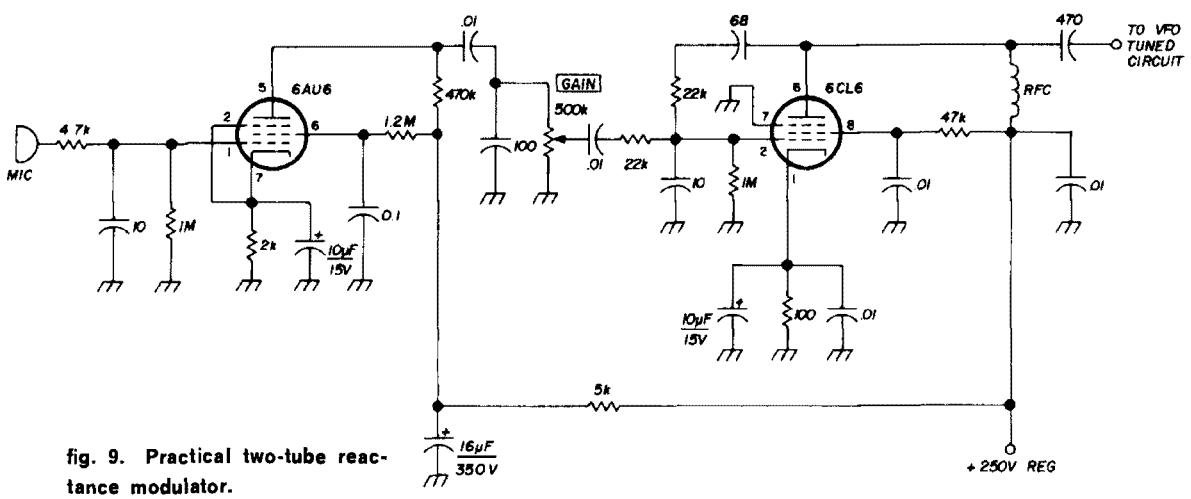


fig. 9. Practical two-tube reactance modulator.

the composite reactance will be varied about its zero-signal value when an audio signal is applied to the modulator. When properly designed and operated, the reactance modulator provides linear frequency modulation and is capable of providing large frequency deviations.

practical reactance modulator circuits

There are many possible configurations for reactance modulator circuits. The principal difference between various arrangements is the type of phase-shifting circuit that is used to produce a grid voltage which is in phase quadrature* with the rf voltage of the modulator plate.

A practical reactance-tube modulator is shown in **fig. 8**. The modulator tube is a high-gain pentode, such as the 6BA6. The plate is coupled through a blocking capacitor to the oscillator circuit; a second capacitor couples the phase-shift network to the modulator grid circuit. If resistor R is made large in comparison to reactance C at the oscillator frequency, the current through R-C will be nearly in phase with the voltage across the tank circuit, and the voltage across C will lag the oscillator tank voltage by almost 90 degrees. The result of the 90 degree lagging voltage on the modulator grid is that its plate current lags the oscillator voltage by 90 degrees, and the reactance tube appears as an inductance in shunt with the oscillator inductance, thus raising the oscillator frequency. Capacitance C in **fig. 8** is often the grid-cathode capacitance of the reactance tube.

Two tubes are used in the practical reactance modulator circuit shown in **fig. 9**. A 6AU6 serves as a high-gain speech amplifier followed by a 6CL6 reactance modulator. The 6CL6's high transconductance permits a large value of lagging current to be drawn during the modulation swing.

A frequency-modulated crystal-controlled oscillator is shown in **fig. 10**. In this circuit a voltage-variable capacitor (**varactor**) is used to alter the resonant frequency

*When two voltages are in quadrature, they are displaced 90° in phase.

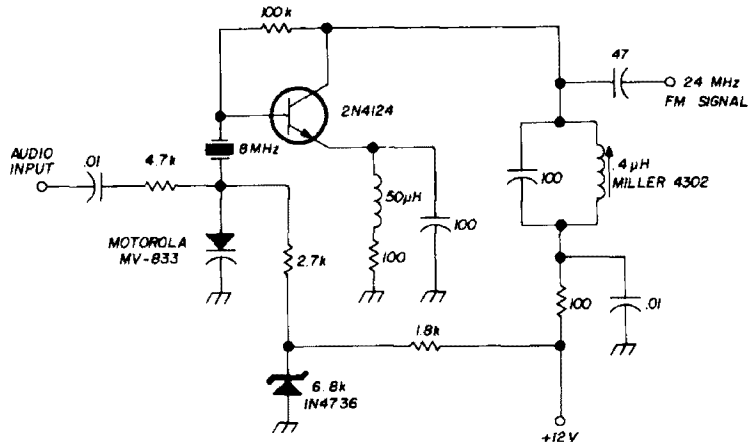


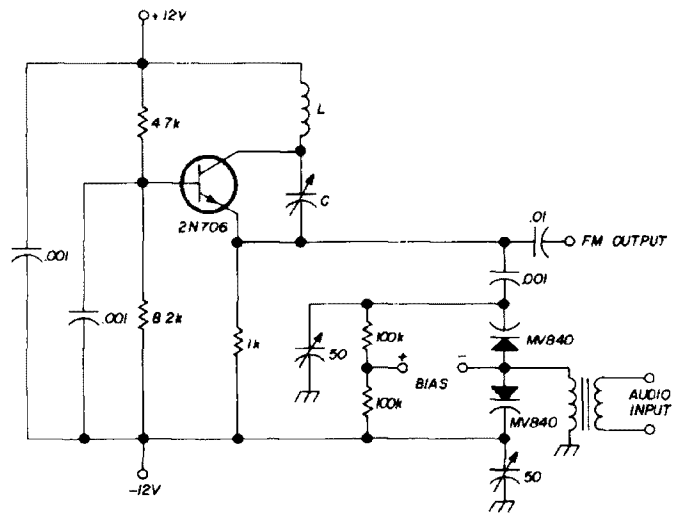
fig. 10. Frequency-modulated crystal oscillator uses a variable-capacitance diode across the crystal.

of the crystal by varying the shunt capacitance to ground. One or two volts of audio are sufficient to provide suitable deviation for transmitter operation at 50 MHz when 8.33-MHz crystals are used. Oscillator output is at 25 MHz.

In the circuit of **fig. 10** the audio voltage driving the varactor appears in the oscillator circuit and produces a small amount of a-m. This is eliminated by the use of two varactors in the back-to-back circuit shown in **fig. 11**. The audio signal is coupled by a high impedance transformer to the diodes through capacitors that are small in comparison with diode capacitance.

The bridge diode circuit presents only a capacitance change, and no component

fig. 11. Back-to-back varactors produce fm without amplitude variations. Modulation transformer should present a high impedance to the diodes.



of bias or drive voltage is presented to the oscillator. Oscillator frequency is determined by L and C, plus the effect of the modulator. Bias voltage is used to adjust the center frequency of the oscillator, providing reverse bias for the diodes to place them in the linear portion of their operating curve. For best results the bias resistors and coupling capacitors should be matched.

determining linearity

It is often desirable to run a static test on a reactance modulator to determine its linearity. A frequency-vs-control voltage curve is plotted on graph paper, and linearity is checked by comparing equal increments in control voltage—both positive and negative—against resulting changes in frequency. A circuit that provides a calibrated negative and positive control voltage is shown in **fig. 12**. The calibrated control voltage may be checked against the frequency deviation produced by the modulator stage and an appropriate graph plotted.

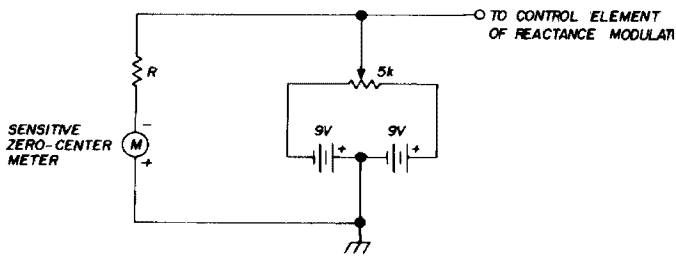
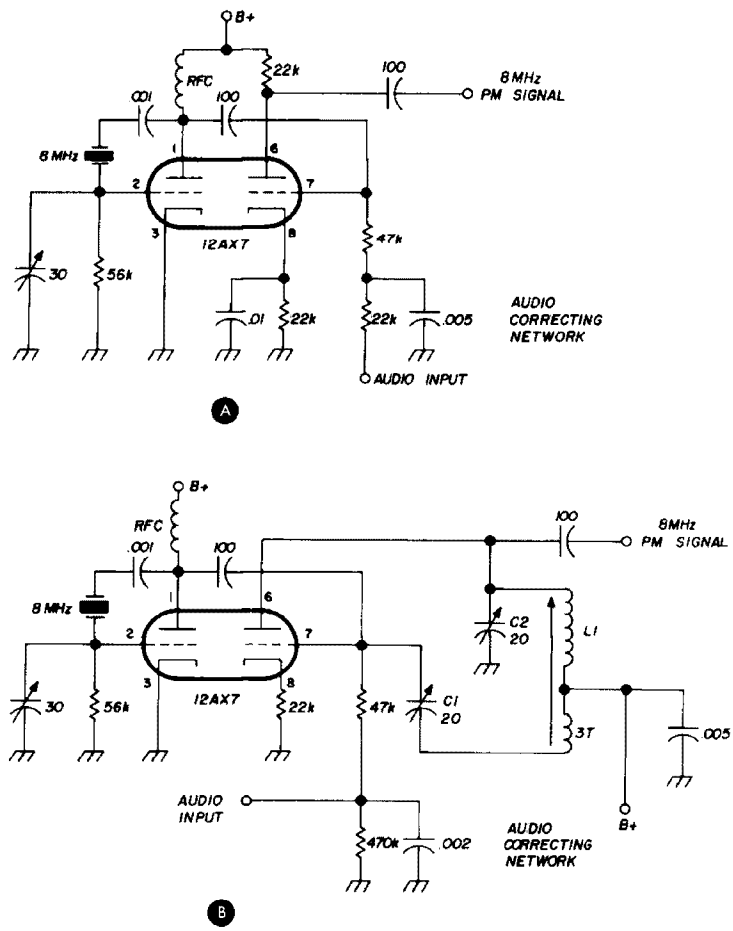


fig. 12. Linearity checking circuit provides calibrated control voltage to the reactance modulator. Resistor R is chosen for full-scale meter deflection when 5k pot is at one end of its range.

fig. 13. Vacuum-tube phase modulators. The circuit in B provides approximately 6 times more deviation than circuit A. Inductor L1 should be approximately 27 μH for resonance at 8 MHz.



generation of pm signals

To generate a phase-modulated signal, the carrier is generated in a constant frequency oscillator, and then passed through a device which introduces phase variations in time with the modulation signal.

A change in the phase of a signal can be produced by passing the signal through a network containing both a resistance and a reactance. If the series combination is considered to be the input, and the output voltage is taken across the resistor, a definite amount of phase shift is introduced, the amount depending upon the frequency of the signal and the ratio of the resistance to the reactance. If the resistance can be varied with an applied audio signal, the phase angle of the output signal changes in direct proportion to the audio signal amplitude and a phase-modulated signal is produced.

Two practical phase modulators are shown in **fig. 13**. Circuit A is an R-C phase-shift network with the phase resistance replaced by the variable plate resistance of a vacuum tube. Since the plate resistance of the triode changes with the applied audio, the phase between the input and the output changes in accord with the audio signal.

In **fig. 13B** phase modulation is adjusted by capacitor C1. Capacitor C2 acts as both a phase angle and magnitude control. Both capacitors are adjusted for maximum modulation. Resonance is established by inductance L1. This circuit provides about six times the deviation of the circuit in **fig. 13A.**

Audio and carrier signals are applied to the control grid. The rf signal reaches the plate circuit by two paths; one through the grid-plate capacitance and the other by the normal amplification action of the tube. When a large, unbypassed cathode resistor is used, tube amplification is greatly reduced, and the two signal voltages across the plate circuit are approximately equal.

The instantaneous plate voltage of the triode is 180° out of phase with the grid voltage, but the feedthrough voltage via the grid-plate capacitance is in phase with the grid voltage. Therefore, the feedthrough is out of phase with the amplified voltage. When a modulating audio voltage is applied to the grid of the modulator, amplification is varied at an audio rate, and the amplified component across the plate circuit likewise varies in amplitude. However, the feedthrough voltage does not change in either amplitude or phase. The **resultant** plate voltage changes in amplitude and phase in accordance with the modulating signal. The small degree of amplitude modulation introduced by this simple modulator can be eliminated in the following stages.

the balanced modulator

Armstrong's original method of generating pm is still widely used, and a variation of the basic circuit is shown in **fig. 14**. Ssb fans will recognize this as a form of balanced modulator which produces a double-sideband, suppressed-carrier signal. The grids of the modulator tubes are fed through an R-C network in the output circuit of the oscillator. Out-of-phase voltages are coupled through small capacitors to a phase-shift network (R1-C3) in the grid circuit of the modulator. This network introduces a 45° phase shift between

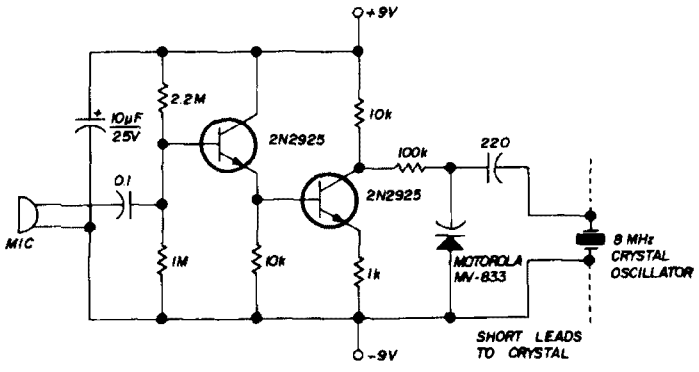


fig. 15. Compact-varactor modulator for low-frequency crystal oscillator.

the grids of the modulator with respect to the reference phase of the oscillator, and 90° with respect to each other.

The suppressor grids are fed by a push-pull audio signal, thus one tube conducts at a time. The combination of balanced modulator action with an rf carrier that

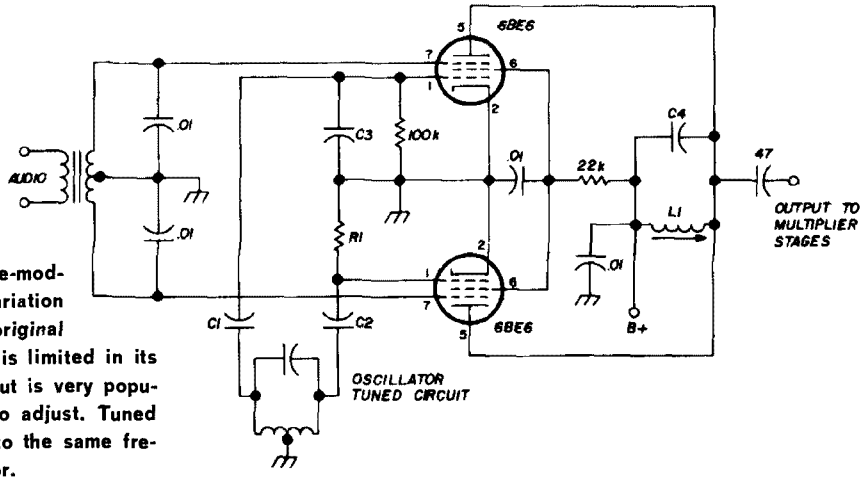


fig. 14. Balanced phase-modulator circuit is a variation of Major Armstrong's original modulator. This circuit is limited in its modulation capability but is very popular because it's easy to adjust. Tuned circuit L1-C4 is tuned to the same frequency as the oscillator.

is shifted in phase by 90° creates a phase-modulated output signal. Thus, amplitude modulation may be converted to phase modulation by removing the carrier, shifting the double sideband 90° and reinserting the carrier.

When the phase-shifted modulation is combined with the carrier, the resultant signal varies in amplitude as well as in frequency. This defect is small when the modulation amplitude is kept small in comparison with the carrier. As a result, this circuit is somewhat limited in modulation capability. It is, nevertheless, used in many types of commercial fm transmitters because of its ease of adjustment.

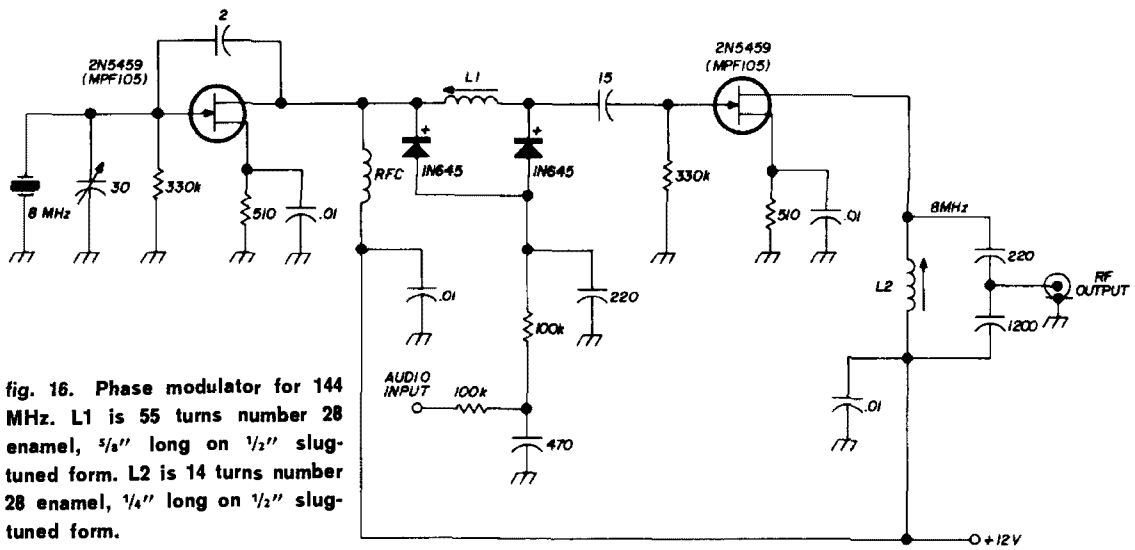


fig. 16. Phase modulator for 144 MHz. L1 is 55 turns number 28 enamel, 5/8" long on 1/2" slug-tuned form. L2 is 14 turns number 28 enamel, 1/4" long on 1/2" slug-tuned form.

solid-state modulators

A solid-state phase modulator that may be used with a crystal oscillator is shown in **fig. 15**. In this circuit a two-stage amplifier places the modulating audio voltage across a varactor which is applied as a shunt capacitance across the crystal. The degree of phase modulation is limited with this approach, and the lowest possible crystal frequency should be used so high deviation ratios can be obtained through frequency multiplication.

A phase modulator designed especially for vhf crystal-controlled fm transmitters is shown in **fig. 16**. The inexpensive field-effect transistor is used in a crystal oscillator operating in the 4- to 8-MHz range. A second fet is used as a phase modulator, with the modulating network in the gate circuit. Two silicon diodes used as varactors across a phasing coil are driven by the modulating voltage. There is a small amount of amplitude modulation with this circuit, but it is "washed out" by the succeeding stages of the fm exciter. The audio system is designed for a high-gain carbon microphone for mobile service. Under these conditions, the rf output of the fet amplifier is about 30 milliwatts.

Bias for the diode modulator is adjusted for maximum deviation with a given amount of audio as monitored in a nearby fm receiver. Channel adjustment is accomplished by the trimmer across the crystal.

conversion of phase modulation to frequency modulation

Phase modulation may be received on an fm receiver with no difficulty, but the modulation will sound harsh and the human voice will sound peaked and tinny. To solve this problem in early designs, the phase modulation was changed so the modulation index decreased in inverse proportion to the modulating frequency. This is accomplished by shaping the response of the transmitter speech amplifier with an audio correction network (**fig. 17**) so that the output voltage is inversely proportional to the modulating frequency. In current commercial practice, flat-response audio with tailored clipping

is fed to the phase modulator in the transmitter (see footnote, page 13); this results in constant peak deviation after modulation, and requires the use of post-equalization at the receiver discriminator.

the fm transmitter

The various direct and indirect methods for producing frequency modulation involve changing either the frequency or the phase of an rf carrier in accordance with the modulating signal. The fm signal is then raised to the operating frequency by passing it through a series of frequency multipliers. When the frequency is multiplied, the frequency deviation is multiplied by a like amount.

Inexpensive, highly stable quartz crystals are available in the 3- to 10-MHz range, and many popular vhf fm transmitters use these crystals, multiplying the crystal frequency by a factor of twelve, eighteen or twenty-four. Some of the earlier military fm gear which used wideband fm in the 20- to 30-MHz region used special low-frequency crystals (type FT-241) that operated in the 250- to 500-kHz region. These crystals allowed sufficient frequency multiplication to obtain the desired degree of deviation at the output frequency.

Generally speaking, the circuitry used in conventional transmitters may be used for fm. In fact, aside from the modulator stage, the circuitry should be familiar to any amateur who has worked with a-m, cw or ssb.

Because the amplitude of an fm signal is constant, the signal may be amplified by nonlinear stages such as doublers and class-C amplifiers without introducing signal distortion. Actually, it is advantageous to pass an fm signal through nonlinear stages, since any vestige of amplitude modulation generated in the reactance modulator may be smoothed out by the inherent limiting action of a class-C amplifier.

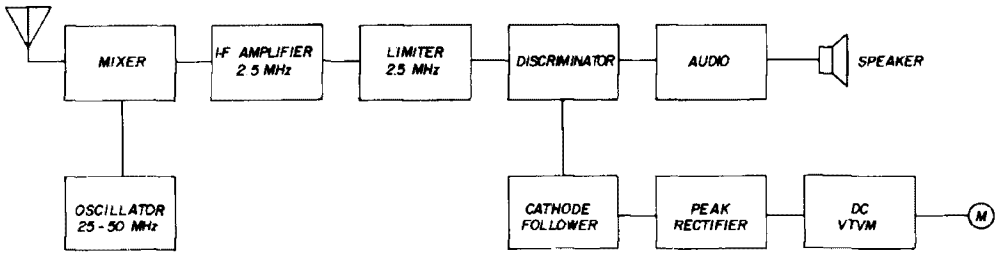


fig. 18. Fm modulation meter depends upon an increase in discriminator output with frequency deviation.

deviation measurement

Deviation of an fm transmitter must be limited to that bandwidth authorized for the service. When a single signal is used to modulate an fm transmitter, the relative amplitudes of the various sideband frequencies vary widely as the deviation is changed as shown in **fig. 4.** Since we know the relationship between the amplitude of the sideband frequencies and carrier to the modulating frequency, a simple method of measuring deviation is possible. This measurement is expressed as the modulation index (m).

The Bessel function plot of an fm signal shows that at certain values of the modulation index, carrier amplitude falls to zero and all the power is contained in the sidebands. By applying a sinusoidal audio signal of known frequency to the

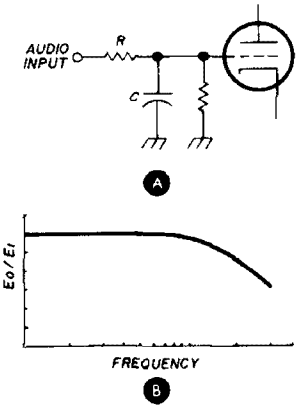


fig. 17. Audio correction network and its effect on frequency response.

transmitter, and increasing the modulation until the amplitude of the fm carrier reaches zero, deviation can be determined. The modulation index at this point may be found from a table, such as accompanies **fig. 4**.

The first point of zero carrier is reached when the modulation index is 2.405 (when the deviation is 2.405 times the modulation frequency). For example, if the transmitter is modulated by a 1000-Hz tone

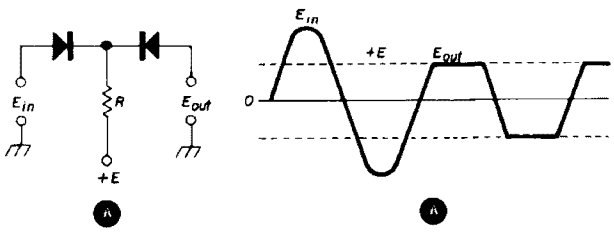


fig. 19. Simple audio clipper and its output waveform. Clipping level is determined by bias voltage E.

and the modulation level increased until the first carrier null is obtained, the deviation will be 2.405 kHz. If the modulation frequency is 2000 Hz, the deviation at the first carrier null is 4.810 kHz. Other carrier nulls occur when the modulation index is 5.52, 8.65, 11.79, etc.

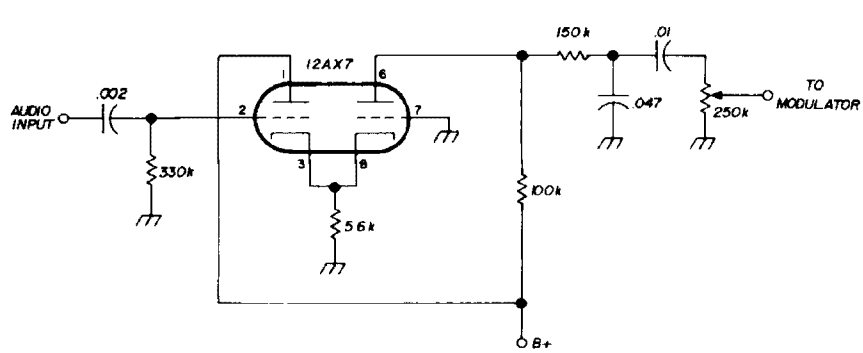
Carrier nulls may be accurately noted with a communication receiver that has a selective i-f system incorporating a crystal or mechanical filter. The output of the fm transmitter is loosely coupled to the front end of the receiver, and the unmodulated carrier is monitored with the bfo turned on and adjusted to a pleasant note, such as 800 Hz. The modulating signal is turned on, and the frequency deviation is varied by increasing the amplitude of the modulating signal until the 800-Hz tone first disappears.

If an fm transmitter has a maximum frequency deviation of 15 kHz, the modulating frequency that will produce carrier disappearance at an index of 2.405 is 6237 Hz ($f = 15{,}000/2.405 = 6237$). An audio signal of this frequency, when applied to the fm modulator and gradually increased in amplitude, will extinguish the carrier. In the measurements setup this adjustment produces a steady lessening of the beat note until it becomes inaudible.

Since 6237 Hz is above the audio range of most communications systems, it may be desirable to use the third carrier null at 8.65. For a maximum deviation of 15 kHz, the test modulating signal would be 1734 Hz (15,000/8.65 = 1734). This test frequency is better for amateur systems since it falls within the desired speech range. For this test, the amplitude of the 1734 Hz tone is increased until the beat note disappears for the third time.

A second method of monitoring deviation is the fm modulation meter. This device consists of a simple fm receiver equipped with a vtvm calibrated in kilohertz of deviation (**fig. 18**). The instrument depends upon an increase in discriminator output as frequency deviation increases. Commercial versions of this instrument will measure modulation of an fm transmitter operating in the range from 25 to 500 MHz. Harmonics of the local oscillator are used at the higher frequencies.

fig. 20. Cathode-coupled twin-triode speech clipper.



modulation limiting

Frequency deviation in an fm transmitter can be controlled with a circuit that keeps the audio signal level within prescribed limits. Simple audio clipping circuits may be used, as well as more complex deviation control circuits that are designed for fm transmission. The audio clipping circuit shown in **fig. 19** uses two back-to-back diodes which clip each half of the audio cycle at a preset level. The clipping level is determined by the bias voltage E. Normally the clipper is followed by a low-pass filter which removes the higher harmonics created by clipping action, thus greatly attenuating sidebands which could cause adjacent channel interference.

A twin-triode cathode-coupled clipping circuit (**fig. 20**) is used in some commercial fm transmitters. When the grid signal at the first triode swings too far positive, the common cathode bias resistor cuts off the second triode; on an excessively large negative driving signal, the first triode section is cut off because of the combined effect of the signal and cathode bias. Therefore, both positive and negative signal peaks are evenly clipped.

fm reception

The fm receiver used by Major Armstrong consisted of several heavy cabinets of equipment that stretched the full length of his laboratory. In 30 years this formidable pile of gear has shrunk to the battery operated, transistorized fm import selling for under ten dollars at the corner drug store.

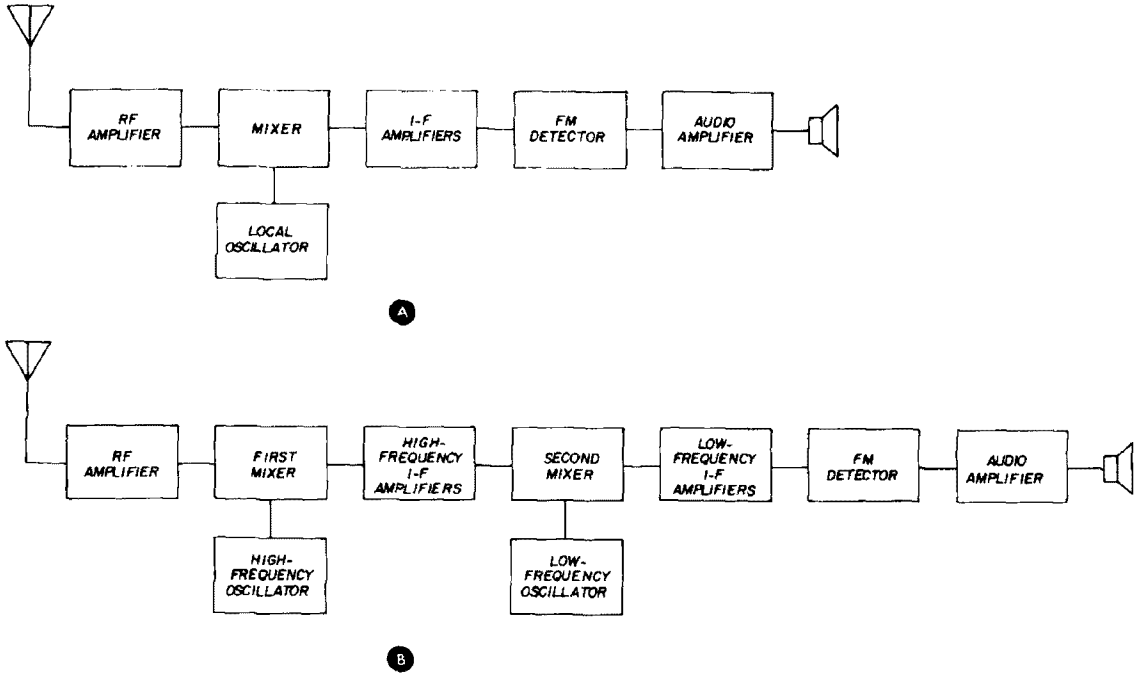


fig. 21. Single- and -double-conversion fm receivers.

An effective fm receiver should be insensitive to changes in signal amplitude and sensitive only to frequency changes in the received signal. By using slope detection, a conventional a-m receiver may be used to receive fm[10] although performance will be much poorer than with a receiver designed for this service.

receiver requirements

There are three main requirements for a practical fm receiver:

1. Bandwidth sufficient to pass the desired fm signal while rejecting adjacent interfering signals.

2. Detector for converting frequency into audible amplitude changes.

3. Limiting systems to eliminate signal amplitude variations before the detector.

A block diagram of a typical fm receiver is shown in **fig. 21A.**

Most vhf fm receivers are double conversion receivers such as **fig. 21B**. When good selectivity is necessary, a low intermediate frequency is desirable; this, however, degrades image signal rejection. Similarly, if good image rejection is required, a high i-f should be used, but this is not compatible with good adjacent channel rejection. These difficulties may be overcome by combining the advantages of low and high i-f amplification in the double-conversion arrangement. First the received signal is mixed with a local oscillator signal to produce a high intermediate frequency; a second local oscillator signal is then mixed with the high i-f signal to produce a low frequency i-f signal.

Low noise rf amplifiers and mixers in fm receivers are the same as their counterparts in a-m receivers. Since most vhf fm receivers work on crystal-controlled channels, the local oscillators may also be crystal controlled. Channel switching is accomplished with either mechanical or electrical switching systems.

The f-m detector requires signals on the order of several volts for proper operation. Since the front end works in the microvolt region, the i-f amplifier must perform most of the voltage amplification. The i-f system must provide sufficient selectivity to discriminate against stations operating on adjacent channels, and still have sufficiently broad response that the outer sidebands of the desired fm signal are not distorted.

To obtain the required selectivity consistent with the proper passband, special bandpass i-f filters are used, such as mechanical or crystal filters, or even L-C filters (Motorola Permakay filter). If the transmitter deviation is changed, the filter may be replaced with one having a passband that matches that of the transmitter.

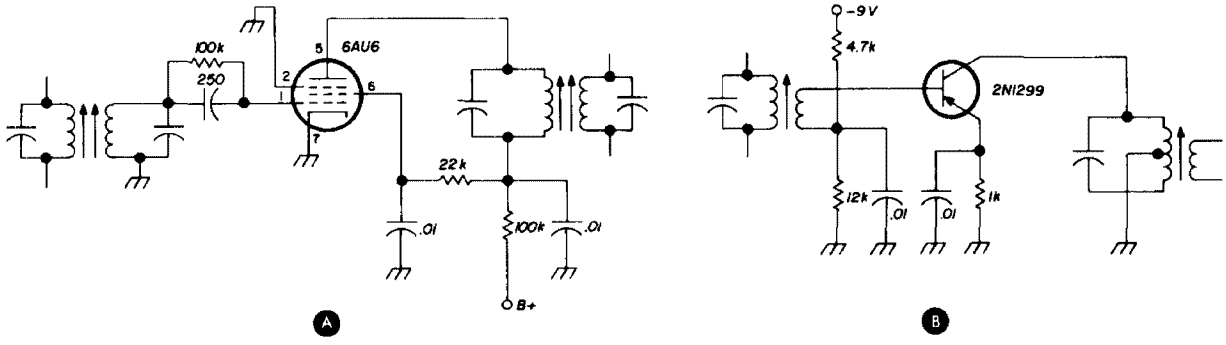


fig. 22. Typical i-f limiter circuits.

i-f limiting

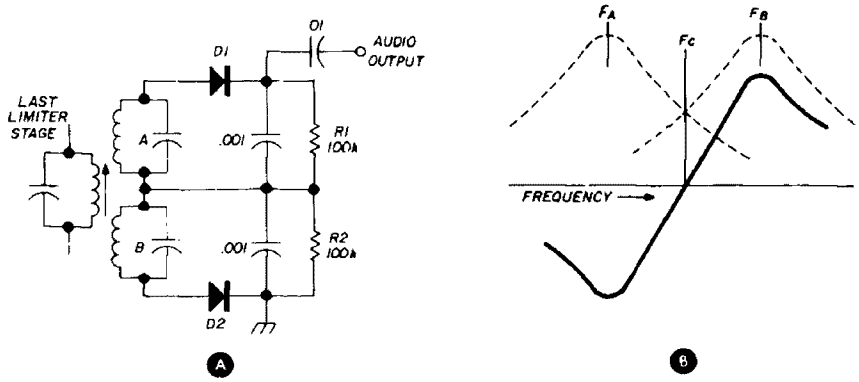
Most fm detectors are sensitive to changes in the amplitude of the received signal so they must be preceded by a limiter that removes all amplitude fluctuations from the signal. The limiting circuit may be an i-f amplifier stage having very low plate voltage and grid-resistor bias so that it overloads easily. Thus, when a certain signal level is reached, further increases in input signal produce no change in the output. If sufficient gain exists ahead of such a stage, variations in amplitude of the received signal are removed. Although a limiter may provide some amplification before the saturation point is reached, its main purpose is to limit the amplitude variations caused by fading and noise.

The limiter circuits shown in **fig. 22** are similar to conventional i-f amplifiers ex-

cept for the bias system. The R-C network produces a bias voltage that is equal to the peak rectified voltage between grid and cathode or between emitter and base. Short amplitude disturbances such as noise are longer than the time constant of the R-C circuit and are clipped off. Longer signal variations such as fading appear in the output. To take care of the slower fading, it is common practice to follow the first short-time-constant limiter with another limiter with a longer time constant.

that is tuned to a frequency that differs somewhat from the average frequency of the fm carrier as shown in **fig. 23**. As the carrier frequency fluctuates, the current in the detuned circuit varies, increasing as the impressed frequency approaches the resonant frequency of the circuit, and decreasing as the impressed frequency departs from the resonant frequency. The output is an a-m wave that may be demodulated by an ordinary diode detector. Since the side of the resonance curve is not linear, only a small

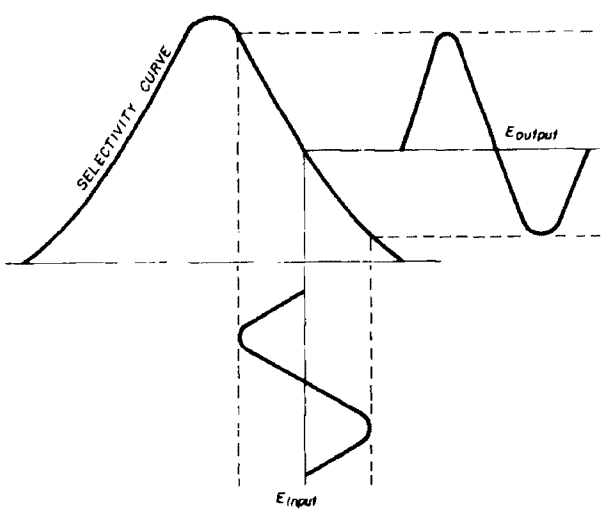
fig. 24. Double-tuned discriminator and its a-m output craracteristic. Curves shown by dotted lines are undesirable responses of this circuit.



fm detectors

A device that converts frequency-modulated signals into audio signals is called a discriminator. The simplest form of discriminator is an ordinary tuned circuit

fig. 23. Slope detection using the response curve of an ordinary tuned circuit.



portion of the resonance curve is useful for linear conversion of frequency variations to amplitude variations.[10] Also, a receiver used in this manner is vulnerable to noise as well as interference from a-m signals appearing at the peak of the resonance curve and fm signals on the opposite slope.

double-tuned discriminators

The linearity can be greatly improved by using two stagger-tuned circuits instead of one and choosing the difference between the two outputs.[11] A double-tuned discriminator circuit is shown in **fig. 24A.** In this circuit the input tuned circuit is resonated to the resting fm carrier frequency, and the resonant frequencies of tuned circuits A and B are spaced slightly more than maximum transmitter deviation. The audio output from this circuit is practically free of distortion because of the linear operating characteristic about the resting carrier frequency.

However, the response curve in **fig. 24B** shows that reception is possible at three points, corresponding to each outer portion of the resonant curves plus the desired linear operating region in the center.

The output from each tuned circuit is applied to a diode detector. When the fm signal is at its resting frequency, the voltages across load resistors R1 and R2 are equal and of opposite phase, and the output voltage is zero. When the fm carrier goes higher than the center frequency, more voltage is developed across one tuned circuit than the other; when the frequency swings lower than the center frequency, the voltage developed across the other tuned circuit is greater. By careful adjustment of the circuit constants the rectified audio voltage is a linear function of the impressed frequency.

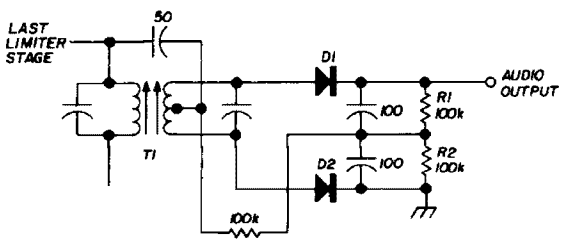


fig. 25. Foster-Seeley discriminator; output voltage vs frequency is the same as fig. 24.

Another discriminator circuit is shown in **fig. 25**. It can be shown that this circuit is essentially a double-tuned pair[12] and that the output vs frequency curve of **fig. 24B** also applies to this circuit.

Circuit operation is based on the phase relationships in a transformer with a tuned secondary. At the resonant frequency of the secondary the rf voltage across each half of the winding is 90° out of phase with the primary; the voltages are 180° out of phase with each other since the winding is center tapped. When the received signal is at resonance, equal rf voltages are applied to each diode; the rectified voltages across the diode load resistors are equal but of opposite polarity so the net voltage output is zero.

When the fm carrier shifts frequency, one diode conducts more than the other because of disturbed circuit balance. If, for example, D1 conducts more than D2, the voltage drop across R1 would be greater than the drop across R2 and the output terminal would be positive. If the carrier shifts in the other direction, D2 conducts more than D1, the voltage drop across R2 increases, that across R1 decreases, and the output terminal is negative. Thus, when the carrier deviates above and below its resting frequency, an audio voltage is developed at the output terminal—the same frequency as the original modulation and proportional to deviation.

ratio detector

A disadvantage of the discriminator circuit of **fig. 25** is that its output is affected by amplitude variations at the input as well as frequency variations. The ratio detector in **fig. 26** was developed as an fm detector that responds only to input frequency variations because the output depends upon the ratio of the currents through the diodes rather than the voltage difference between them. Since the ratio detector is insensitive to a-m, it is not necessary to use a limiter ahead of it.

In the ratio detector in **fig. 26**, the large

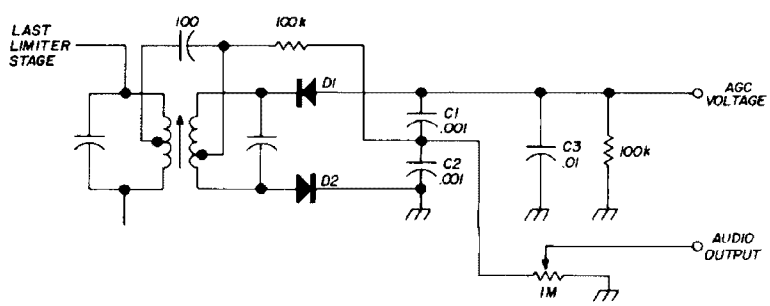
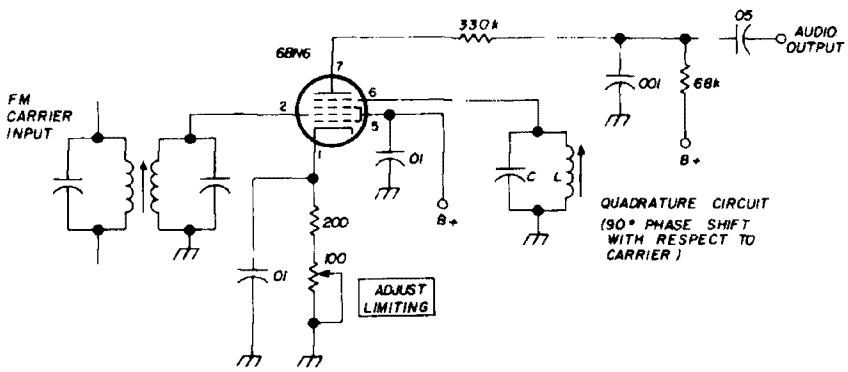


fig. 26. Ratio detector is relatively insensitive to amplitude modulation.

value of C3 tends to maintain the voltage across C1 and C2 at a constant level. As with the two discriminator circuits discussed earlier, the relative average currents through the diodes vary with input frequency. The voltages across C1 and C2 vary correspondingly, one increasing as the other decreases, while their sum remains constant. If the two diode circuits are carefully balanced, the output voltage is practically independent of any amplitude variations on the incoming signal.

on it. The square waves from the limiting amplifier are rectified by the diodes to negative pulses which charge the output capacitor C. As the frequency increases, the average rate of charge increases; when the frequency decreases, so does the voltage across the capacitor. The charge on the capacitor depends upon the deviation of the signal from the resting frequency. The shunt resistor lowers the time constant of the circuit so the stored charge can change fast enough to reproduce an audio signal.

fig. 28. Gated-beam discriminator uses specially designed vacuum tube. Quadrature circuit is tuned to the same frequency as the fm carrier.



At the carrier resting frequency, C1 and C2 are charged to a constant value. When the frequency decreases, C2 acquires a greater charge than C1; when the frequency increases, C2 is charged less than C1, although the sum of the charges remains the same. The audio output at the junction of C1 and C2 follows the frequency deviation.

cycle-counting detector

An fm detector that uses no tuned circuits is shown in **fig. 27.** This cycle-counting detector operates with a resistance-coupled i-f amplifier and responds directly to the number of cycles impressed up-

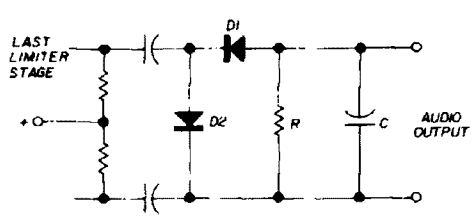


fig. 27. Cycle-counting detector charges capacitor C at rate proportional to modulation frequency.

Finally, mention should be made of the 6BN6 gated-beam detector shown in **fig. 28.** This fm demodulator is a special single-tube circuit that features excellent sensitivity and high output. With one or two volts on the input, incoming signals need little amplitude to cut the tube off for negative peaks; similarly, for positive peaks of the incoming signal, saturation is reached. Limiting action may be adjusted by a variable bias resistor in the cathode circuit.

Discriminator action in the gated-beam tube takes place as a result of signal phase shift between two control grids which are 90° out of phase with each other by virtue of a quadrature circuit connected between them. When the quadrature circuit is properly tuned, the electron beam is modulated in proportion to the modulation on the incoming signal.

squelch circuits

Squelch circuits are used to mute the audio when no signals are present. In a

high-gain receiver, speaker noise can be very annoying to operators who must monitor a channel for long periods. When the receiver is squelched, no background noise is heard; when an rf signal comes on, squelch is turned off and the audio system becomes operative.

A simple squelch circuit is shown in **fig. 29**. The squelch voltage is derived from

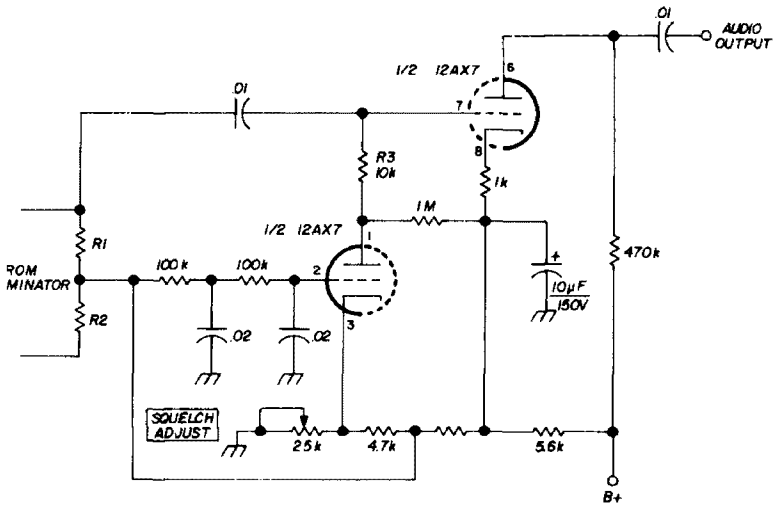


fig. 29. Discriminator derived squelch circuit.

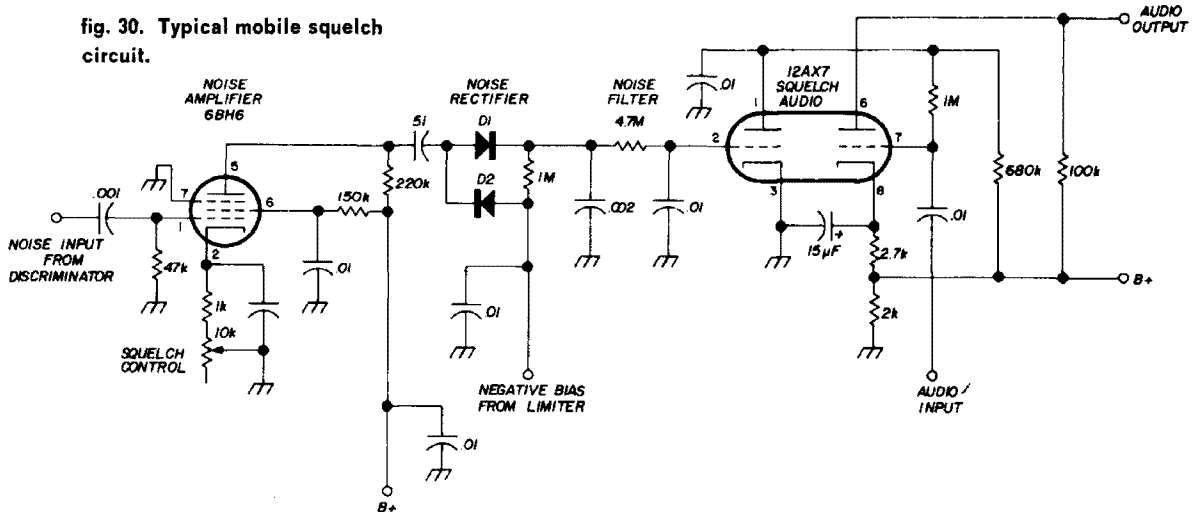
the discriminator circuit. The center of the discriminator load resistors develops a negative voltage when a signal is applied. This voltage is passed through an R-C network and is sufficient to bias the squelch tube to cutoff and permit a signal to pass through the audio amplifier. Without an input signal, a positive bias (applied to the junction of R1 and R2) is greater than the bias applied to the cathode of the squelch tube; this causes the tube to conduct, dropping the plate voltage and cutting off the audio tube through resistor R3.

A more complex squelch circuit is shown in **fig. 30.** This circuit consists of three stages; a noise amplifier, a noise rectifier and a squelch control tube. In the absence of signal, the limiter stages of the fm receiver pass a continuous and constant level of noise voltage to the discriminator. Noise higher than the voice frequencies is amplified and passed to the noise rectifier developing positive voltage across the load resistor.

A negative bias voltage from the grid circuit of the second limiter is applied to the anode end of the load resistor. The sum of the negative limiter voltage and the positive rectified noise voltage appears across the load resistor. When a signal is received, the signal voltage overrides the noise voltage, reducing the positive noise voltage at the noise rectifier.

The dc amplifier is biased near cutoff, the audio stage is biased to normal operating conditions, and squelch is removed. The variable squelch level control is set to limit the gain of the noise amplifier tube, thus limiting the noise voltage to a value that will just bias the audio stage to cutoff. Hence, when even a weak carrier signal is received which quiets the noise, the squelch is opened.

fig. 30. Typical mobile squelch circuit.



transistorized squelch

The transistorized squelch circuit shown in **fig. 31** consists of a noise amplifier Q1, noise detector D1, and a squelch gate (switching circuit) Q2. In the absence of an rf carrier, noise output from the discriminator is amplified by Q1. The ac output of Q1 is rectified and applied as bias to the base of Q2. With no noise signal applied, Q2 is cut off and Q3 is conducting; thus, the desired audio signals are amplified by Q3 and applied to the audio driver stage Q4.

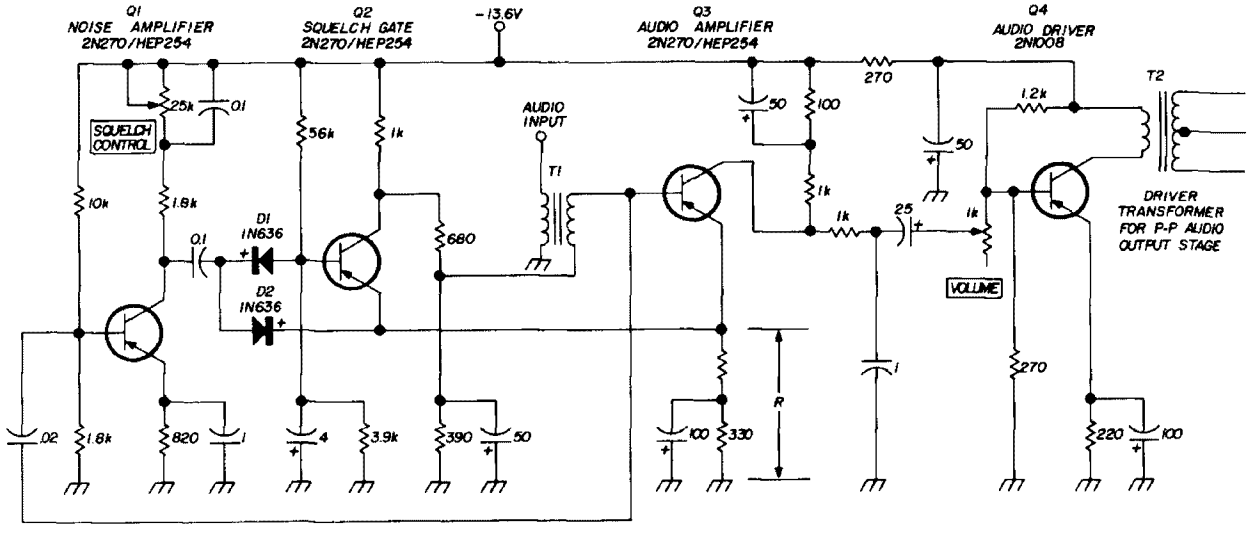


fig. 31. Transistorized squelch circuit.

In the absence of carrier, noise at frequencies at 4000 Hz or higher is applied to noise amplifier Q1, rectified and applied to the base of squelch-gate transistor Q2. When this voltage reaches a predetermined level (***threshold***) *determined by the squelch control*, the squelch-gate transistor conducts. Since the squelch-gate and preamplifier emitters are returned to ground through a common emitter resistance (R), conduction of the squelch gate results in a more negative voltage at the emitter of the audio amplifier, Q3. This increased negative voltage cuts off the audio stage and no signal reaches the audio driver.

Most of the objectionable noise is in the frequency range above 4000 Hz, but the desired audio is in the 300- to 3000-Hz range. The squelch circuit is designed so that signals above 4000 Hz open the squelch gate and signals below 4000 Hz do not. This is accomplished in the input circuit to the noise amplifier. The time constant is such that signals below 4000 Hz are greatly attenuated and do not provide sufficient rectified voltage to overcome the squelch threshold voltage at the base of the squelch-gate transistor.

Some squelch circuits substitute a relay in place of the blocking tube or transistor, with the relay contacts breaking the speaker circuit or activating auxiliary control circuits when the fm carrier appears.

automatic frequency control

Some fm receivers are equipped with an automatic-frequency-control (afc) circuit which adjusts the frequency of the mixer oscillator *to compensate for variations in the* received signal frequency or frequency drift in the oscillator.

In an afc system, a discriminator is used to provide a dc voltage which is proportional to the frequency of the received signal. The potential at the output of the discriminator is used to control a reactance modulator shunted across the oscillator tuning circuit. Circuit constants are chosen so the tuned circuit varies the oscillator frequency to cancel out drift. The response of the system must be sufficiently slow so that variations in carrier frequency caused by modu-

lation have no effect on the afc; only slow variations in the average frequency are able to operate the reactance tube (**fig. 32**).

fm repeaters

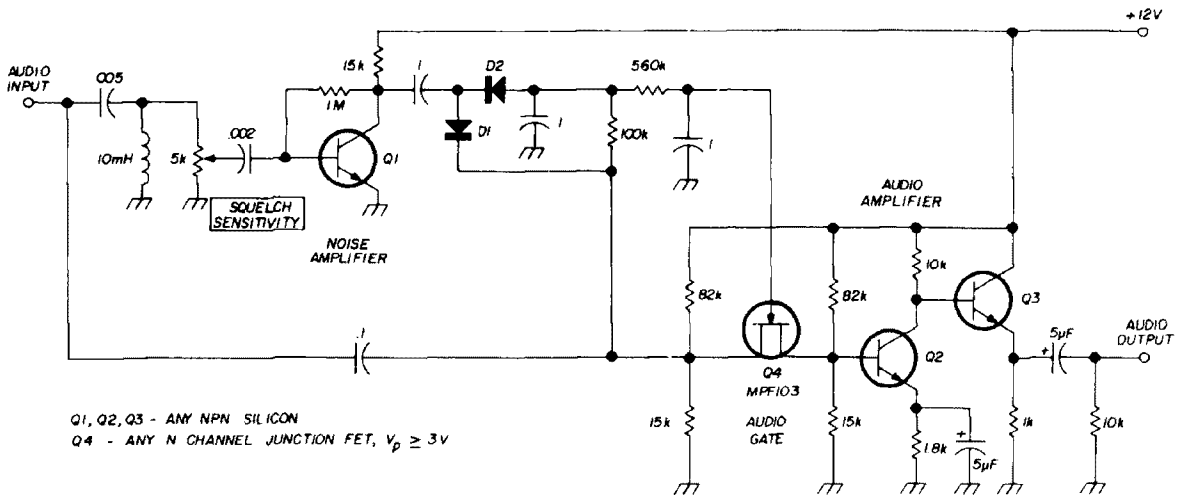
Vhf radio transmission is essentially short range, limited to line of sight unless sophisticated equipment is available to take advantage of unusual modes of propagation. A two-way radio relay station or repeater may be used to extend the communication range between two or more fm stations. Various types of repeaters are in use in the United States and Canada, their operation and mode depending upon the requirements of the circuit they cover.

A repeater receives on one frequency and transmits on another. It may use separate antennas or a single antenna and diplexer so both reception and transmission can occur simultaneously. The user's transmitter is on the input frequency of the repeater, while his receiver is tuned to the output. The remote base (**fig. 34**) is a form of repeater whose location has a height or tactical advantage.

An unattended repeater may be operated by remote control under the provisions of Section 97.43 of the FCC Rules; control may be by wire, or by radio on an amateur frequency above 220 MHz. In some instances, remote repeaters serve on common frequencies so that individual groups operating their own installations may cross-communicate.

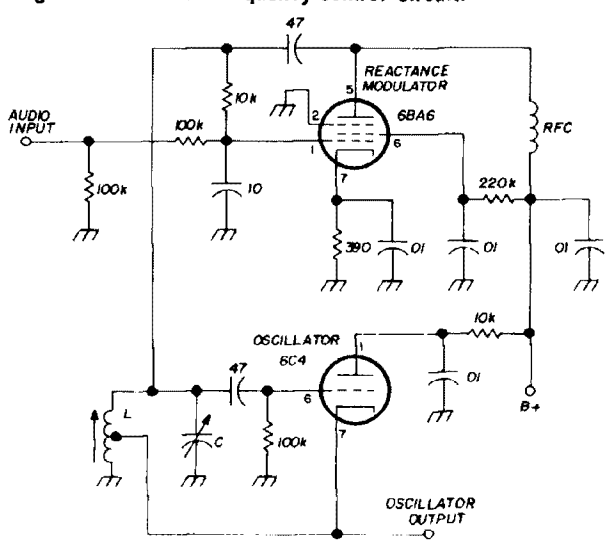
The repeater receiver is normally in a standby condition, as is the transmitter section. Upon remote command a carrier-operated relay is allowed to place the repeater on the air, with receiver audio fed through a coupler to the transmitter modulator. After transmissions are complete the relay opens and the repeater is removed from the air.

Simplex operation refers to communication between individual units operating on a common transmit and receive frequency, which may, in turn, be interfaced with repeater operation, using either a local or re-



Solid-state squelch circuit suggested by K6HWJ uses an n-channel fet as a series audio gate.

fig. 32. Automatic-frequency-control circuit.



mote base. While most repeaters are limited to a single communication channel, multiplex operation, or simultaneous transmission of two or more signals on a common carrier, is sometimes employed by means of narrow band filter techniques. For example, simultaneous voice and RTTY transmissions may occur on a single fm channel. It is also

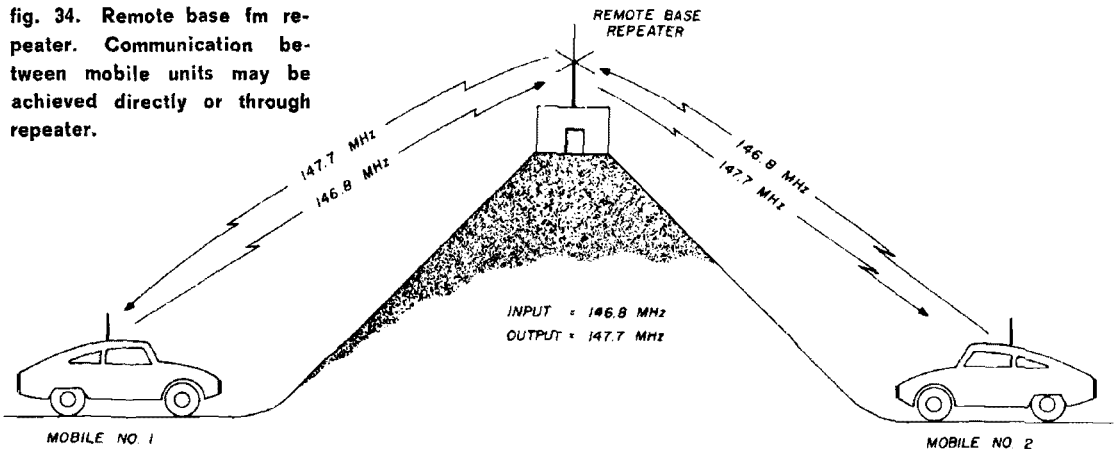


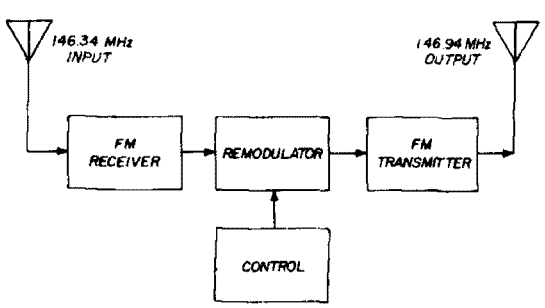
fig. 34. Remote base fm repeater. Communication between mobile units may be achieved directly or through repeater.

possible to insert traffic at a repeater for transmission in either or both directions. **Duplex** (simultaneous two-way) transmissions through the repeater may be achieved in many cases.

vhf amateur fm operating standards

Generally speaking, vhf amateur fm techniques are based upon the channel concept. Transmitters and receivers are mainly crystal controlled on a given channel and random tuning techniques common to the lower frequency amateur bands are absent. Channel spacing on the 6-meter band is usually 40 kHz, beginning with 52.56 MHz, in addition to the national calling channel at 52.525 MHz. Some narrow-deviation fm is used on the West Coast on 51.0, 51.1, 51.2 and 51.3 MHz, with additional channels near these frequencies for limited-access repeaters. Fm channels on the 2-meter band are usually 60 kHz apart, starting at 146.040 MHz, including the national calling frequency at 146.94 MHz. A spot frequency of 29.6 MHz is often used for fm operation on 10 meters, with channel spacing of 40 kHz starting at 29.040 MHz. On the 420-MHz band, channel spacing is 50 kHz. Standard deviation for all bands is 15 kHz, with some trend towards narrower deviation of 5 kHz. Antenna polarization is vertical. In some areas where both wide- and narrow-band equipment is in use, a compromise deviation of 7.5 kHz peak is used.

fig. 33. Block diagram of a typical vhf fm repeater.



getting started

Vhf fm is an increasingly popular facet of amateur radio. The contrast in operating techniques from the QRM-filled lower frequency bands appeals to many operators. A big advantage of fm is the "quiet intercom" style of operation. Crystal-controlled transmitters and receivers eliminate tuning and missed call letters, fading is conspicuously absent, and interference is minimum.

Radio clubs and technical groups find fm perfect for club communication, and semi-private fm nets are springing up in the larger cities, as are "open" nets. Several nationwide channels have been established so that

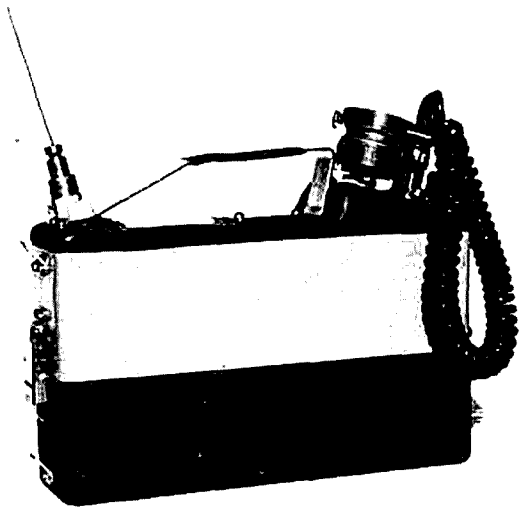
fm'ers from one area can quickly establish communication in another area when mobiling across the country.

Second-hand fm equipment is available from many sources at prices ranging from ten dollars for "junkers" up to $100 or so. A good mobile two-way fm installation should probably be available for under $70, depending upon power, model and condition of the gear. Motorola, General Electric and RCA are some of the more popular makes of equipment, as circuit diagrams and replacement parts are readily available.

Some imported equipment for 2- and 6-meter fm is starting to show up on the American market; and other U.S. brands such as Link, DuMont, Kaar, Aerotron and Bendix are common also. Conversion of commercial gear for amateur use entails the purchase of new crystals and retuning of various stages to the amateur bands. Some vhf commercial gear (for the 148- to 170-MHz range) requires padding capacitors in the rf stages.

A gold mine of information on the latest in amateur fm techniques and equipment may be found in the monthly magazine *FM*[13] For additional background on fm techniques, the reader is referred to references 14 to 22.

This battery-operated fm handie-talkie provides one watt output on two meters and is typical of surplus gear used on the amateur bands.



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ham radio

using integrated circuits with single-polarity power supplies

This clears up the question of power-supply connections to linear IC's and presents some hints on lead dress and bypassing

John J. Schultz, W2EEY, 40 Rossie Street, Mystic, Connecticut 06355

Although an IC may specify a dual-polarity power supply, a single-polarity supply can usually be used. This applies to almost all IC's but should be of particular interest to those who read my article in the May 1968 issue of **ham radio**.[1]

The article showed the uses for an IC operational amplifier in amateur radio circuit applications. Apparently, however, some readers have been confused as to how the power-supply connections to the IC should be handled. Since similar confusion can develop when using an IC—not just the one shown in the article—I thought it would be worthwhile to explain in some detail how proper power supply connections to an IC can be made.

basic mistake

Perhaps the simplest mistake was to misunderstand the power-supply voltage requirements. For instance, in **fig. 1,** which is a simple wideband rf amplifier, some builders connected a 6-volt battery between terminals 4 and 6, and others connected a 12-volt battery between the same terminals. Both connections are wrong. The IC won't function and will probably be ruined by the 12-volt battery. Note there are three reference points for the power supply connections: terminals 1 and 3, 4 and 6. Terminal 4 must be 6 volts negative with respect to terminals 1 and 3. Thus, two 6-volt batteries are required, or a power supply that will deliver a plus and minus 6-volt output. This basic dual-voltage requirement is common to most operational amplifier IC's.

a dual-polarity supply

There should be no problem in obtaining the dual-polarity voltages. The simple circuit

of **fig. 2** can be used for light loads. An additional rectifier/filter can also be included in an existing power supply to obtain the voltage (of either polarity). However, situations

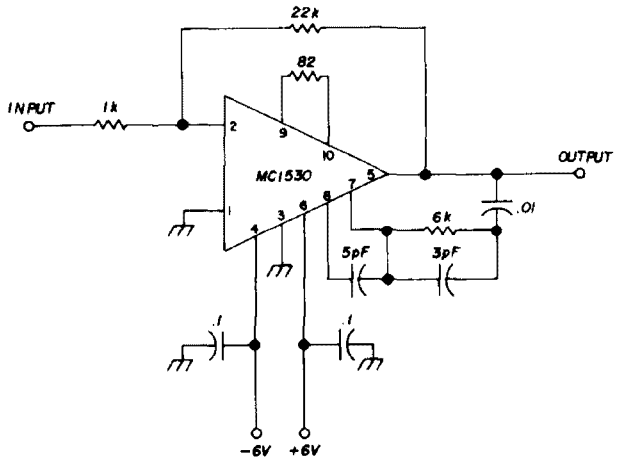


fig. 1. Wideband IC amplifier is normally wired for use with a dual-polarity power supply.

exist where you might want to use only a single-voltage supply, either because only one battery source is available (such as in an automobile), or it's too complicated to modify an existing supply that has an exceptionally well-filtered output of given polarity. Therefore, it's worthwhile to see how IC's can be modified to operate from a single-voltage power supply.

operation from single-voltage power supply

As mentioned for the IC shown in **fig. 1,** three reference points exist for supplying power: −6, 0, and +6 volts (or simply −V, 0, and +V on other IC's). Whatever power supply is used, these reference levels must be maintained. One possibility for using a single-voltage supply is to make the reference levels 0, +V, and ++V. Thus a 12-volt supply with a 6-volt tap could be used to power the IC shown in **fig. 1.**

The most important point to remember, however, is that the former zero reference (terminals 1 and 3 in **fig. 1**) now requires a +V level. You cannot simply connect terminal 4 in **fig. 1** to ground, place a 12-volt source on terminal 6, and leave all other connections alone. Terminal 1 of the input differential amplifier in the IC is normally at ground potential when using dual-supply voltages and must be biased to one-half the total single-supply voltage.

ic's with external and internal grounds

Some IC's have an external ground terminal, as shown in **fig. 3A,** while others have only an internal ground reference point, as shown in the equivalent circuit of **fig. 4A.** This doesn't change the basic condition for properly biasing the input terminal when an IC designed for a dual-voltage-polarity power supply is used with a single-polarity supply. An example of how each IC can be biased is shown in **figs. 3B** and **4B.** The circuit in **fig. 3B** shows only the dc power connections to the IC and is quite straightforward.

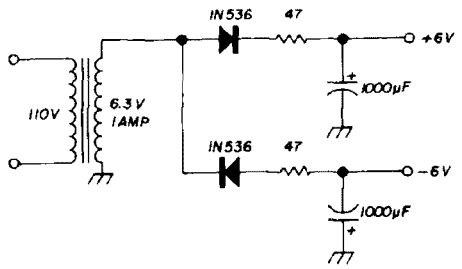


fig. 2. Simple power supply that delivers dual-polarity output. With some audio circuits it may be necessary to add another 47-ohm resistor/1000 μF capacitor in each output lead to reduce hum to a very low level.

Some IC's of the type shown in **fig. 4** have a terminal marked "offset adjustment" which, for most amplifier uses, is connected to the +V terminal of the IC. Isolating resistors R1 and R2 in **figs. 3B** and **4B** ensure that the IC input impedance is not degraded by the power supply connection. Resistor R1 can usually be from 50k to 1000k ohms. R2 may or may not be required, depending on what use was made of this differential input terminal in the original circuit. If the terminal in the original was grounded, R2 will probably not be needed, and terminal 1 in

figs. 3B and **4B** can be connected directly to +V.

If the original circuit had terminal 1 connected in some sort of feedback or dual-input function, then R2 must be used. Its value can be best determined by experimentation, so that the original circuit function isn't disturbed. A practical rule, however, is to make R2's value equal to about the parallel combination of R1 and the feedback resistor used in a specific circuit (the 22k-ohm resistor between terminals 5 and 2 in **fig. 1**).

some samples

Probably the best way to clarify the principles just mentioned is to illustrate their application to a specific circuit. Suppose you want to operate the wideband amplifier shown in **fig. 1** from a single-polarity power supply. **Fig. 5** shows two ways of accomplishing this.

In **fig. 5A** a zener voltage divider is used to obtain 6 and 12 volts from the power supply. Terminal 6 of the IC goes to the 12-volt point. The terminal 4 dc-reference shown in **fig. 1** is now connected directly to ground. The 100k-ohm resistor between terminals 1 and 2 performs the isolating function of R1 as described for **fig. 3B.**

Fig. 5B shows almost exactly the same connections, except that a simple resistive voltage divider, instead of a zener-diode divider, is used to obtain the 6-volt level.

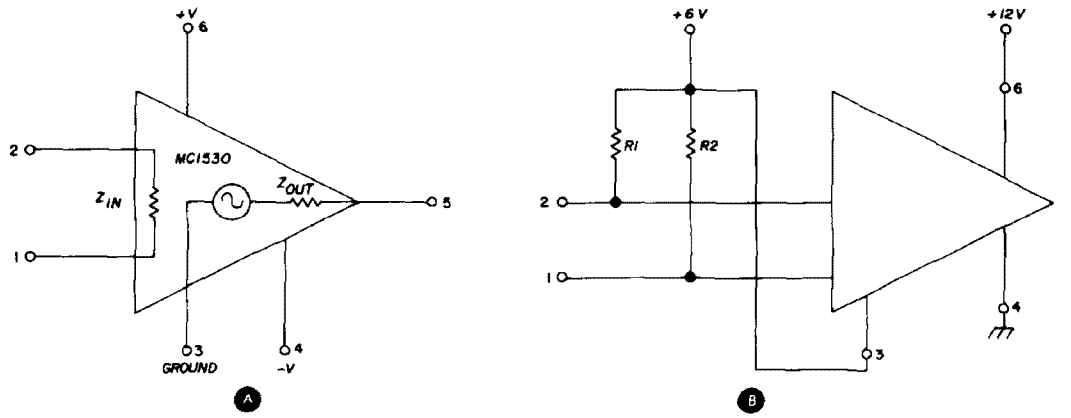


fig. 3. Equivalent circuit and signal-polarity supply connection for an IC having an external ground terminal. Signal connections are not shown.

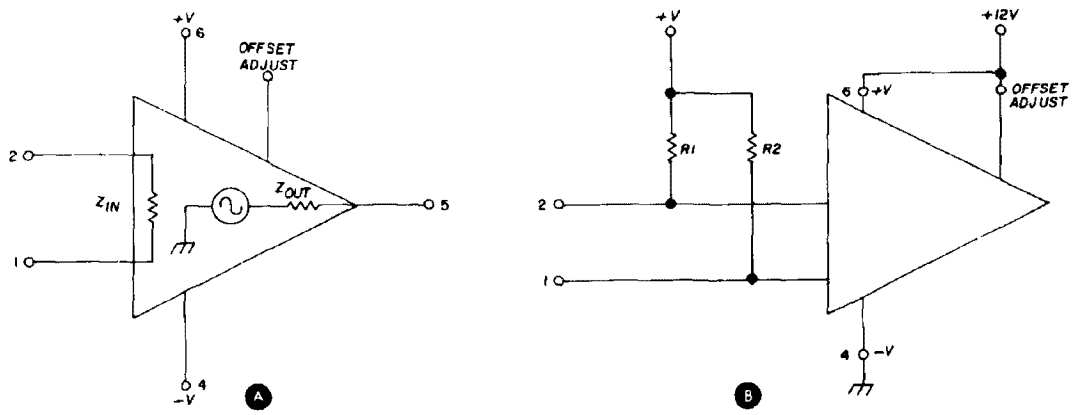


fig. 4. Equivalent circuit and single-polarity supply connection for an IC with an internal *ground* terminal. Terminal marked "offset adjustment" may not be present on all operation-amplifier ICs.

Either circuit can be used, depending on the components available and the regulation of the basic power supply.

Note that either circuit places a dc-bias voltage on the input and output terminals of the IC, which was not present before. Therefore, if the original circuit used direct coupling to the input and output stages, a dc-blocking capacitor will be required in the input and output leads. Normally, this won't alter circuit performance. The only exception would be if the IC were used as a dc amplifier. In this case, additional circuitry may be necessary, or it may not be feasible to operate the IC from a single-polarity power supply.

an exception

Almost all IC's require the techniques described. However, if a unit designed for a dual-polarity power supply is to be used properly with a single-polarity supply, you'll find that manufacturers are designing their IC's to be used with different power supplies. The IC internal circuits are arranged to avoid the necessity for biasing the IC input terminals for different power supplies.

Fig. 6 illustrates such an IC: the Motorola

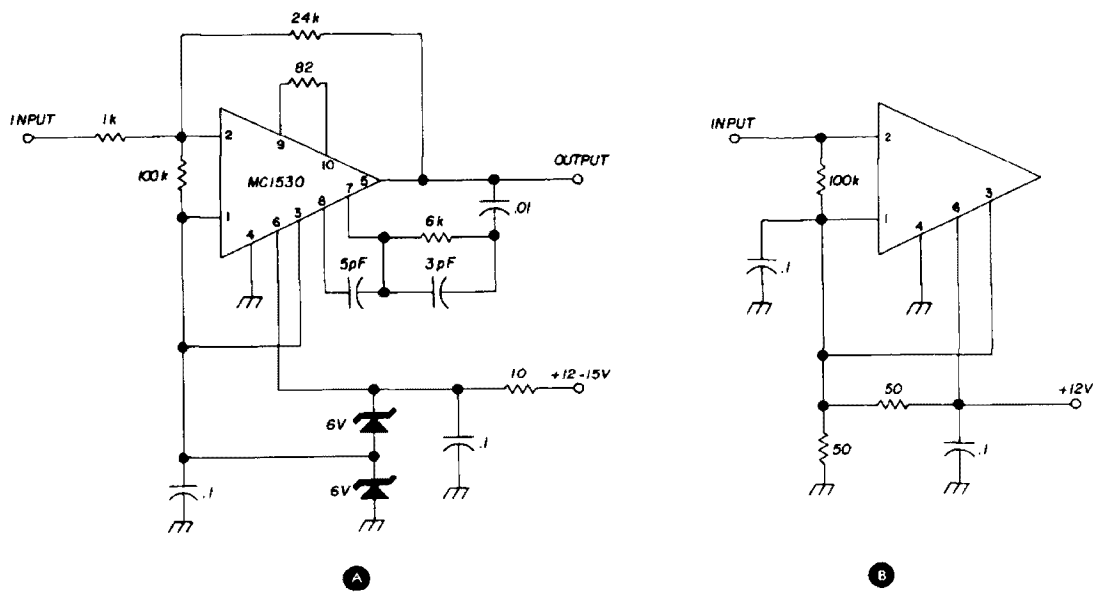


fig. 5. Two methods of providing a voltage divider. Complete circuit is shown in A; only dc-power connections are shown in B.

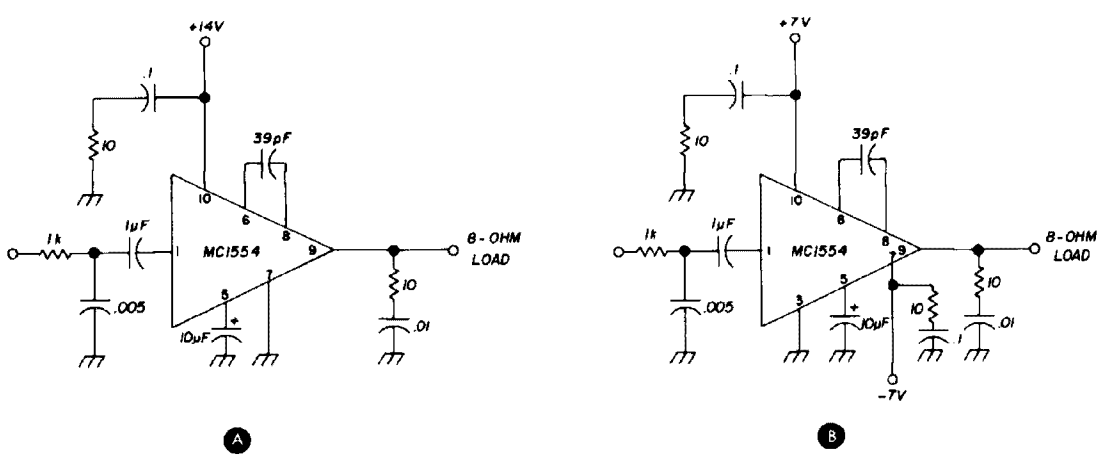


fig. 6. The Motorola MC1554 (HEP 593) audio power amplifier IC is especially designed to simplify operation from either a single- or dual-polarity supply (A and B respectively).

MC1554, which is a complete 1-watt audio power amplifier. **Fig. 6A** shows IC connections using a single-polarity power supply, and **fig. 6A** those using a dual-polarity supply. The connections are similar and involve no modification to the input circuit. Terminal 10 (V+) is connected to the positive power-supply terminal in both cases. Terminal 7 (V−) is connected either to ground or to the negative power-supply terminal. Terminal 3, called the bias-reference terminal to avoid confusion with the V-terminal (which may also be connected to ground), is either unused or grounded as shown.

decoupling and suppression

Care must be taken in any amplifier to prevent oscillation and instability due to signal coupling via power supply or other leads. The requirements are even more stringent for IC's because (1) the IC is a concentrated package of high gain capability, and (2) it has an extremely wide frequency response. An audio-amplifier IC, for instance, may have a response up to several hundred kHz. Therefore, feedback capacitances or coupling circuits small enough not to affect a tube or transistor audio amplifer may well cause an IC amplifier to oscillate. These oscillations may not be heard because of their high frequency, but nonetheless can cause the IC to heat up and possibly be destroyed.

The 10-ohm resistor and the 0.1-μF capacitors shown in **fig. 5A,** the 0.1-μF capacitors shown in **fig. 5B,** and the series 10-ohm and 0.1- or 0.01-μF networks shown in **fig. 6** are examples of decoupling or bypassing components. They should never be eliminated to simplify construction. These components are shown at various places in the schematics, but in practice they should be mounted as close as possible to the IC terminal with which they are associated. All lead lengths should be kept as short as possible to avoid the buildup of lead inductance, which can cause instability.

reference

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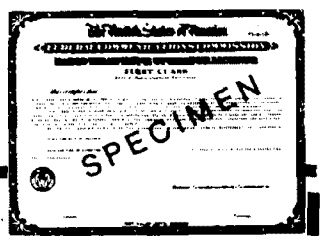
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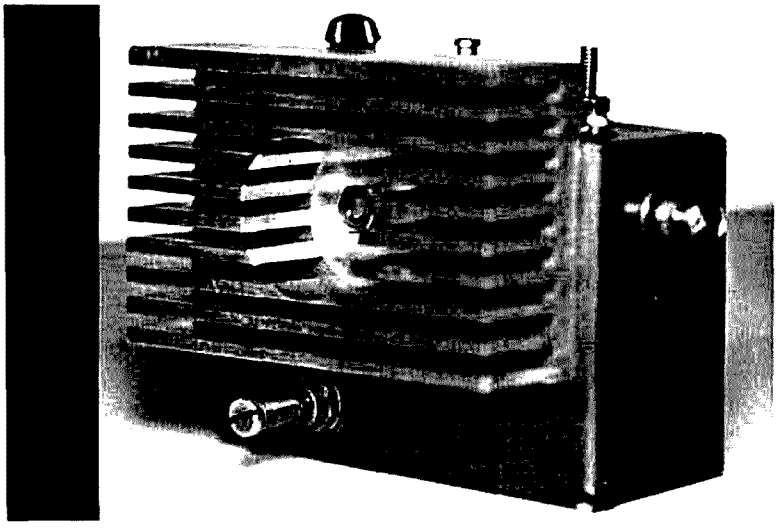
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Introducing the varactor as an efficient microwave harmonic generator for transmitting use

Joseph J. Moraski, K4SUM, Charles E. Spitz, W4API,
1420 S. Randolph Street, Arlington, Virginia 22204

Of the many pursuits available to experimenters in amateur radio, perhaps none offers more challenge and personal satisfaction than working in the frequency region above one gigahertz. It's challenging because tubes and components obey physical laws in a different manner than at the lower frequencies. Personal satisfaction occurs when you've mastered the techniques and special discipline required to make microwave circuits perform properly (or at all).

The objectives of this article are to encourage more amateur activity in our microwave bands and to show how we resolved one of the many problems one encounters at the "last frontier"—obtaining useful drive power in a frequency tripler at 1296 MHz through the use of a simple homemade cavity.

some background information

An excellent discussion of the behavior of grid-controlled power tubes at the higher frequencies is contained in reference 1. One of the things the article points out is that, as frequency increases, the internal structure of the tube becomes an appreciable part of the resonant circuit. Driv-

ing voltage decreases because of the reactance presented by tube lead inductance. Because of this and other peculiarities, tubes operating at uhf and beyond are inefficient as frequency multipliers, at least in amateur work.

the varactor

Solid-state technology has developed a device that seems, presently, to be the only answer for obtaining usable amounts of power with reasonable efficiency in multiplier circuits at the higher frequencies. The device is known as a varactor, or voltage-variable capacitance diode.

By definition a diode is a two-terminal p-n junction. Normally it operates as a rectifier (forward conduction), or as a voltage regulator (zener) conducting in the reverse direction. True, the varactor **is** a two-terminal p-n junction, but it's not correct to call it a diode per se. As with all p-n junctions, the varactor has a junction capacitance. This is what makes the varactor work as a frequency multiplier. The junction capacitance varies with applied voltage, which results in harmonic production.

A very readable treatment of varactor principles is contained in reference 2. Equations are derived showing the relationship of junction capacitance and applied voltage. The end result is a voltage across the varactor containing two dc components and a second component rich in harmonic content.

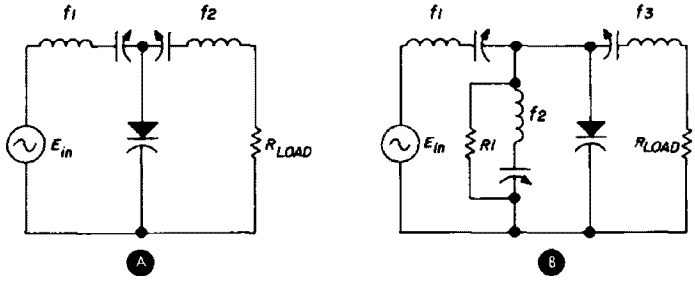
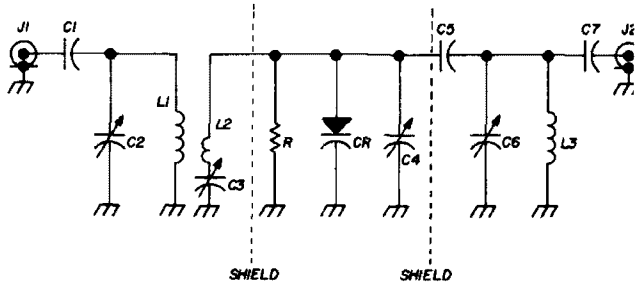


fig. 1. The varactor as a frequency multiplier. In A f1 is tuned to the fundamental and f2 to its second harmonic. Part B shows a tripler circuit with f2 tuned to the second harmonic (the so-called idler). Series circuit f3 is tuned to the third harmonic of f1. Bias voltage is developed across R1 when the varactor is driven into conduction on peaks of the input voltage.

higher-order harmonic generation

In varactor application if frequencies higher than the second harmonic are desired, idler circuits are used. These are tuned circuits that pass or attenuate harmonic currents to obtain the desired output frequency. **Fig. 1** shows the development of a frequency multiplier varactor for generating third-order harmonics. The tuned circuits are affected by the varactor capacitance, which in turn varies as a function of the applied voltage. In a harmonic generator this voltage varies sinusoidally, so the **average** varactor capacitance is used in the design. The capacitance varies with signal power, and some circuit detuning occurs if there is a substantial change in input power. The effect is undesirable, especially in a-m ap-



C1, C5, C7	**Brass or copper disc, 0.4 inch diameter**
C2, C6	**Brass or copper rod, 10/32 thread**
C3	**0.6 — 30 pF piston (Johanson JMC 1901*)**
C4	**0.6 — 5 pF piston (Johanson JMC-1375)**
L1	**Brass or copper tubing, 5/16 inch diameter, 2.55 inches long**
L2	**No. 14 copper wire, 3½ inches long**
L3	**Brass or copper tubing, 5/16 inch diameter, 1.1 inch long**
J1, J2	**Single-hole-mount female BNC connectors**
R1	**100k to 1 megohm, ¼ W (see text)**

fig. 2. Schematic of the 1296-MHz tripler. The varactor can be a Motorola 1N5149, Amperex H4A (1N4885), or Microwave Associates MA-4060—all were successfully tried in this circuit.

* Surplus units available from Fertik's Electronics, 5249 "D" Street, Philadelphia, Pennsylvania, 19120. The small ones are 3 for $1.00; larger ones are $.65 each.

plications, but is practically eliminated in step-recovery junction varactors, as discussed below.

junction characteristics

Varactors are of two types, step junction and step recovery junction. In the former, the impurity level is constant in the p-n layers, while in the latter the impurity level is concentrated at the lead contacts and decreases toward the junction. This makes for high resistivity near the junction and low resistivity at the lead contacts.

An important parameter of varactors is the equivalent series resistance, composed of the bulk and contact resistance of the semiconductor material. This resistance is directly related to varactor efficiency, and therefore kept as low as possible. By employing a constantly decreasing impurity profile from lead contact to junction, the average series resistance remains at a desirable low value when a reverse voltage is applied. This is because the depletion layer dissipates the high resistivity region. The average series resistance is therefore kept low.

Why use step recovery junction varactors for harmonic generators? Because they provide higher power and a more linear relationship of power output to power input. What this means in terms of efficiency (power out/power in x 100) is demonstrated in **table 1,** which shows the power linearity characteristic of a typical step-recovery varactor used as a frequency doubler.

The average efficiency of this particular varactor is about 66 percent, which is typical when operating a frequency doubler. However, we were able to obtain yields of 50-percent plus, using similar varactors as **frequency triplers** with output at **1296 MHz.**

Essential to the performance of any rf generator are the tuned circuits. This is especially true at microwave frequencies for the reasons pointed out earlier. The easiest way to solve this problem is to use resonant cavities. If the procedures used in commercial practice are followed, cavities are difficult for the average amateur to build. However cavities need not be round, as in most commercial applications. Frequencies in the 1-GHz region are high enough to permit a small rectangular chassis to be used, with divider strips acting as cavity partitions. Commercial test jigs have been made in a similar manner, but with facilities and materials beyond usual amateur resources.

If you're reasonably adept with hand

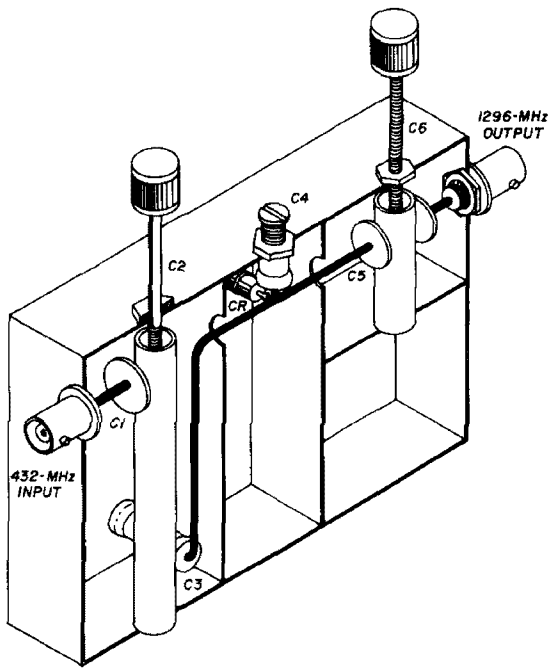


fig. 3. Assembled 1296-MHz tripler.

tools and like to work with soft metal, you should be able to duplicate the tripler described here. With an input of 20 to 25 watts on 432 MHz, we built three units that put out a respectable 10 to 15 watts on 1296 MHz. All have been operating for almost a year. The schematic is shown in **fig. 2.** Capacitors C_1, C_5, and C_7 are single discs; the other capacitor plate is formed by the tubing (L_1 and L_3).

The chassis and cavity partitions were made from 0.032-inch sheet brass. It might be possible to use a commercially available chassis of nearly the same dimensions, such as the Bud CU-341 Converta-box. The partitions could be made and added as shown. However, we haven't tried this, and for best results we recommend using the materials and dimensions shown. The tripler we built is shown in the

photo and in **fig. 3**. **Fig. 4** shows component layout.

In **fig. 3** the varactor stud protrudes through the rear wall where a heat sink was placed over the stud. A solder lug was placed at the varactor base for grounding the bias resistor. The resistor is shunted directly across the varactor.

The brass sheet is cut to the shape of **fig. 5**.* Drill holes to sizes indicated. When

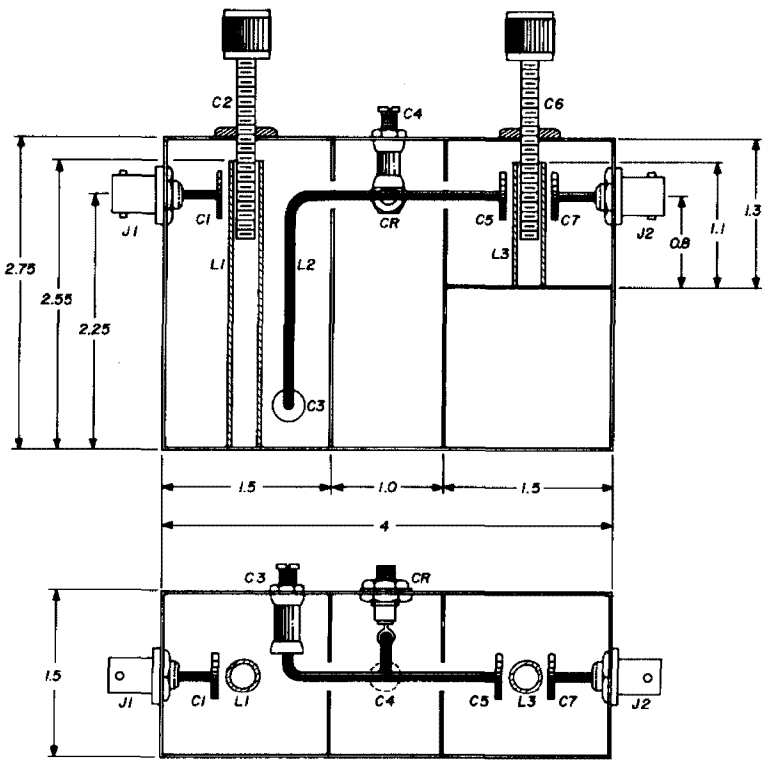


fig. 4. Plan and side views of the 1296-MHz varacter tripler.

the ends are bent, a chassis will be formed 4 inches wide x 2-3/4 inches deep x 1/2 inch high. Solder corners using a heavy iron or torch. Next, cut out the partitions and cover following the scale-size template. Insert partitions C, D and E making sure the holes D and F are drilled before soldering the partitions into place.

The cover is also shown in **fig. 5**. Bend the corners as indicated, solder smoothly, file, and buff. The idea is to get a good fit without warping the chassis.

The 432-MHz and 1296-MHz cavities are capacitively tuned by turning their 10-32 threaded rods (C_2 and C_6) into tubing inductances L_1 and L_3. It is important to maintain constant metal-to-metal contact between the tuning screws and the chassis. A 10-32 brass nut is soldered over holes A. It may be necessary to use a locking nut and spring between the knob and soldered nut if the threads don't mate well enough for continuous contact.

Solder the bases of the brass tubing flush with holes D. Try to make this a tight and straight fit before soldering. Screw the threaded rods through the tubing to ensure centering and alignment, then solder the base. This is where a torch works best. A teflon washer may be forced on the bottom of each threaded rod to prevent shorting or teflon spaghetti may be used.

The next step is to install BNC fittings J_1 and J_2 with their discs, C_1 and C_7. Spacing between the discs and the brass tubing is approximately 1/32 inch. On one model we made, the BNC nuts were soldered to the outside of the chassis so these spacings could be made variable. A short wire lead will be necessary between the disc and BNC inner conductor. This dimension is not given, as the length of the center conductor may vary between different BNC chassis fittings. The critical factor is the spacing of the disc.

Install plunger tuning capacitors C_3 and C_4 with serrated washers under the nuts. Install L_2 in one piece, soldering one end to C_3 and the other to the C_5 disc. Approximate spacing of C_5 from L_3 is about 1/16 inch. Be very careful to apply **minimum heat** or use heat sinks when soldering to the plunger capacitors. Too much heat can crack the glass or warp the glass-to-metal seals so the plungers won't enter the stator cups. Many plunger capacitors are ruined this way. Solder only to lugs or pins provided.

Place a solder lug on the varactor stud, then place the stud through hole I. Fasten with serrated washer and nut. If a heat sink is used, make it secure but not overly

* Full-scale templates available from **ham radio** magazine for $.25.

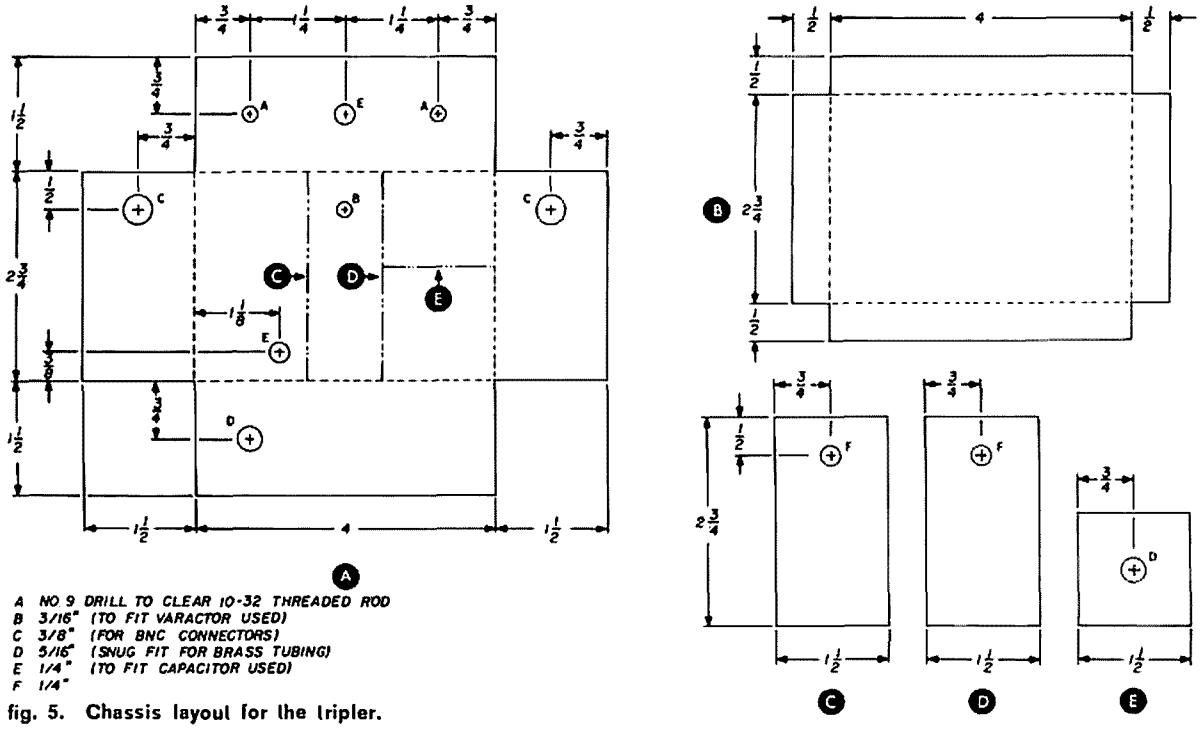


fig. 5. Chassis layout for the tripler.

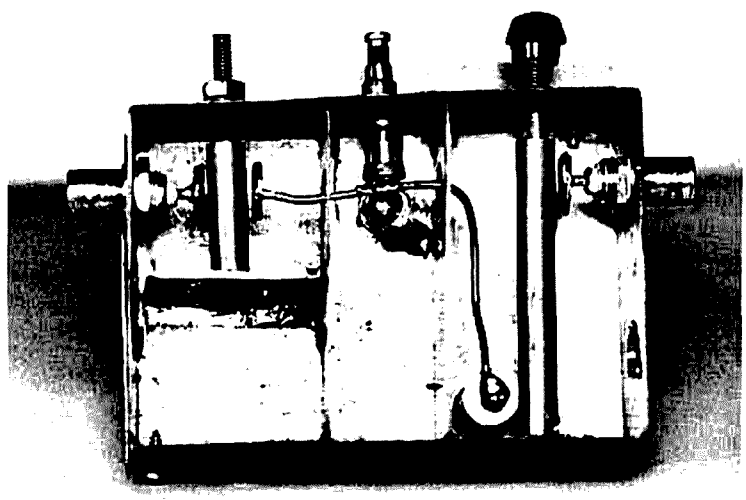
tight until soldering is done, when the nut can be loosened for final placement of the heat sink. Solder the resistor to the lug as reasonably short and straight as possible so it's parallel with the varactor. Join varactor, resistor, and C_4 idler capacitor at a common point on L_2, then solder. Maximum output will be obtained with a 1-megohm resistor; however if the driver is modulated greater linearity will result if 100k is used.

The teflon tubing may be kept from falling out by using Scotch X-1181 copper foil tape with conducting pressure-sensitive adhesive. The tape can also be used instead of finger stock to seal the compartments. "Sticky Fingers," a pressure-sensitive-backed finger stock made by Instrument Specialties Company, of Little Falls, New Jersey was also considered, but the size of the available stock was too large for use with the small cavity.

cool it

Overheating the varactor at this power level can and does occur. The DO-5 case mounting stud, which protrudes through the chassis may be felt for heat. Another sign of overheating is a gradual decrease in output power with no change in the input. The best preventive measure is to use a heat sink and preferably air from a fan. We used some finned aluminum heat sink stock 4 inches long x 2½ inches wide x 1 inch deep. It was a piece of surplus used for power transistors and had one flat side. The varactor stud goes through a hole in the heat sink, then is fastened with its nut and washer to clamp the

Underchassis view of the 1296-MHz tripler. Note lead dress, especially of capacity C5. Some bending might be necessary to optimize output.



varactor, heat sink and chassis together. A Whisper fan furnished more than enough air to keep the varactor cool at extended periods with maximum power.

operation and tuning

The input and output levels were measured and monitored using a Bird Model 43 coaxial line section, panel meter, and 400-1000-MHz element. With this element readings are about 1 dB low on 1296, but the cost was about one-third of an element with greater accuracy. Frequency readings were taken with a surplus F-26/UPR wave trap, which is calibrated from 1.25 to 6.0 GHz. Although the idler frequency is off scale at 864 MHz, it can be read. Since varactors will readily double with high efficiency, care must be taken to minimize idler output while maximizing 1296-MHz output during tuneup. The surplus TS-186 frequency meter, which has a range of 100 MHz through 10,000 MHz, with visual as well as audible output, may also be useful. While all this test gear is not absolutely necessary, it does speed things up for the initial alignment.

If the output is less than at least 7 watts with 20 watts input, and all tuning controls have been optimized, a greater yield may be possible by repositioning the disc capacitors, either in spacing or by a slight upward or downward deflection. However, each change must be accompanied by completely retuning all variable elements for maximum output at the desired frequency. A great deal of time can be saved if you can monitor the input power, output power, and output frequency during tuneup. If you do get 50 percent efficiency, leave it alone!

An input filter at 432 MHz is necessary if you use a stripline varactor tripler, a tube tripler, or grounded-grid stage as a driver, because all these circuits contain harmonic and mixing frequencies that would be seen and used by the 1296-MHz varactor tripler. On the other hand, the tripler output cavity should have about 30 dB rejection of unwanted frequencies when properly adjusted. If the output is fed directly to an antenna, a filter may be called for. If it's used to drive an amplifier with tuned input and output circuits at 1296, one may not be necessary.

table 1. Power conversion in watts of the IN5150 without retuning.

Output	Input
4	4
6	9
8	12
10	16
12	18
14	20

$f_{in} = 0.5$ GHz, $f_{out} = 1$ GHz

The cover should be used at all times. This reduces radiation loss, prevents having to retune when the cover is placed on the chassis, isolates the compartments in the chassis, and avoids a real hazard of biological injury from radiation at these power levels. If screws are used to fasten the cover, sheet-metal screws as small as possible should be used, with two at the top and bottom of the varactor compartment, one at the center of the 432-MHz cavity, and one just below the bottom of the 1296-MHz cavity.

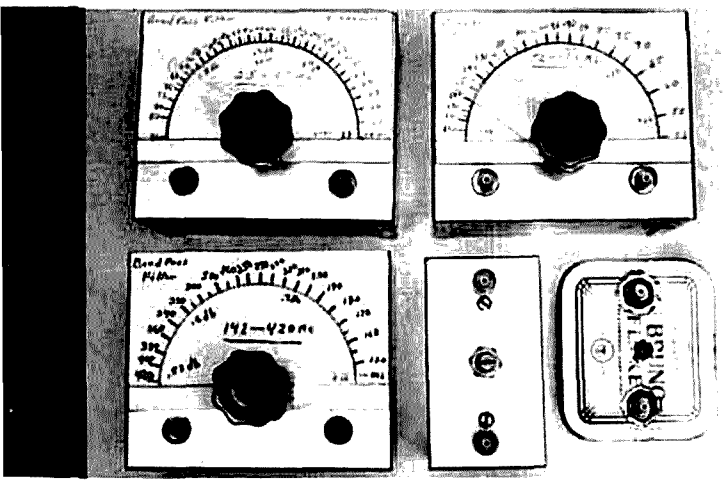
Fairly acceptable modulated signals for local use have been produced where the basic modulated transmitter was on 144 MHz and passed through two triplers. With the varactor input reduced so that a cool ICAS output of 7.5 watts was maintained on 1296 MHz, a pair of 7289's (improved 2C39A, 3CX100A5) as an amplifier produced 40 watts output with a plate voltage of 800 and plate current of 200 mA. There was no sign of creeping or overheating.

Construction and operation of the unit is simpler than it might appear from the description given, but we felt these details would be helpful since most amateur data on the subject seemed meager.

references

1. R. I. Sutherland, W6UOV, "Vhf/uhf Effects in Gridded Tubes," *ham radio*, January, 1969, p. 8.
2. "The Semiconductor Data Book," Motorola, Inc., Semiconductor Products Division, 2nd Edition, p. 16-89.

ham radio



tunable bandpass filters for 25 to 2500 mhz

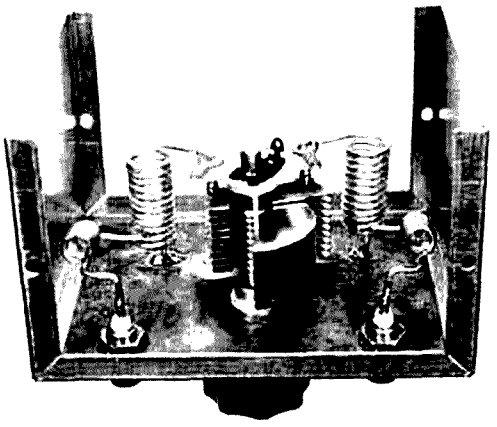
Here are some useful additions to your vhf and uhf test equipment

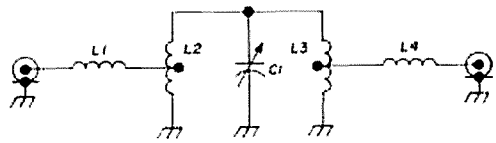
Del Crowell, K6RIL, 1674 Morgan Street, Mountain View, California 94040

For all their advantages, solid-state devices are not without their problems, especially when used in amplifier and multiplier circuits in the vhf and uhf regions. If you have used these devices in such applications, you've probably run into the same problems I have—is the output at the desired frequency?

The bandpass filters and accessories described here are great additions to your test equipment. They will allow you to build and adjust many circuits that would otherwise be

Construction of a bandpass filter that covers from 25 to 85 MHz.





Filter	Frequency Range (MHz)	L1, L4	L2, L3	C1
A	25 to 85	6 turns no. 14 ¼" diameter, ⅝" long	12 turns no. 14, ½" diameter, 1½" long; tapped 1½ turns from ground	9—140 pF (Hammarlund MC140M)
B	52 to 190	4 turns no. 16 ¼" diameter, ½" long	4 turns no. 16, ½" diameter, ⅝" long; tapped 1½ turns from ground	7.7—100 pF (Hammarlund MC100M)
C	142 to 420	2 turns no. 16 3/16" diameter ¼" long	3½ turns no. 14, 3/16" diameter, ⅜" long; tapped 2 turns from ground	6.3—50 pF (Hammarlund MC50M)
D	230 to 1350	see fig. 3	see fig. 3	0.7—30 pF piston
E	2000 to 2500	see fig. 5	see fig. 5	see fig. 5

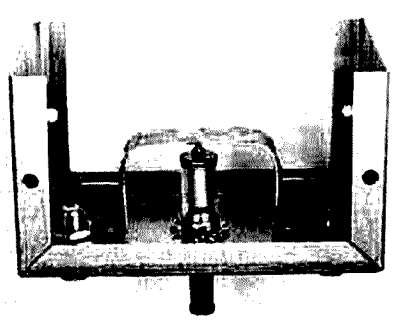
fig. 1. Bandpass filters cover the frequency range from 25 to 1350 MHz and 2.0 to 2.5 GHz.

almost impossible. An accurately calibrated, general coverage receiver with a panadaptor will give good results when making circuit adjustments using these filters.

description

Filters A through D in **fig. 1** use a half-wave resonant circuit with series matching coils. This helps maintain a constant impedance and provides a balanced circuit with good operating Q. The circuit diagram and coil table show the L-C relationships. Bandwidth is primarily governed by the amount of loading, which is determined by the tap point on L2 and L3. L1 and L2 are used to maintain a more constant loading with frequency; the tap points given in the table are optimum for general-purpose use. Filters with a tuning range of three to four octaves can be constructed using this principle.

Bandpass filter for 230 to 1350 MHz uses a brass-strap inductor.



The 2.4 GHz bandpass filter is built around an old tobacco tin.



construction

Bandpass filters covering 25 to 420 MHz are basically the same, with only L and C changes required to cover the desired ranges. The filters can be duplicated by referring to **fig. 1** and the photos. Filters A, B and C are constructed using a brass plate

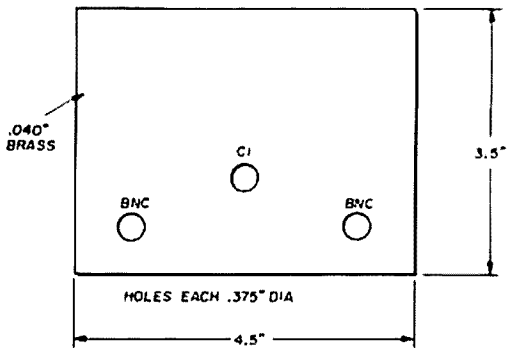


fig. 2. Brass mounting plate for filters A, B and C.

(see **fig. 2**) to mount the components. Coils are soldered directly to this plate. The assembly is then mounted inside an LMB779 5x4x3-inch chassis box.

The filters can be calibrated with a grid-dip oscillator; however, a good signal generator would be more accurate, of course. Referring to the circuit in **fig. 1,** the 3-dB bandwidth and total insertion loss can be varied by changing the points on L2 and L3. Tapping down on the coils gives greater selectivity but more insertion loss.

Filter D uses the same principle as A, B and C, except a strap inductor is used in place of coils L2 and L3 (see **fig. 3**). An expensive capacitor was used in the model shown in the photo, but the simple screw-and-disc capacitor shown in **fig. 4** will do the job nicely. A low minimum capacity must be used to reach above 1296 MHz.

Filter E is mounted in a discarded pipe tobacco can. I drilled holes in the ends for the input-output connectors and soldered a coupling wire from the connector to the

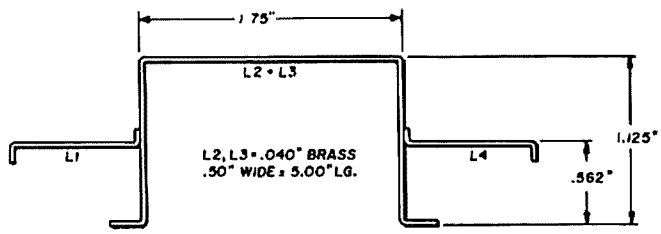


fig. 3. Inductor detail for filter D.

fig. 4. C1 for filter D is made by soldering a disc to a brass screw about 1½" long; an aircraft-type plate nut is used as a bearing.

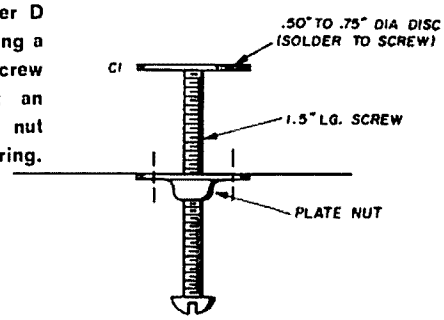
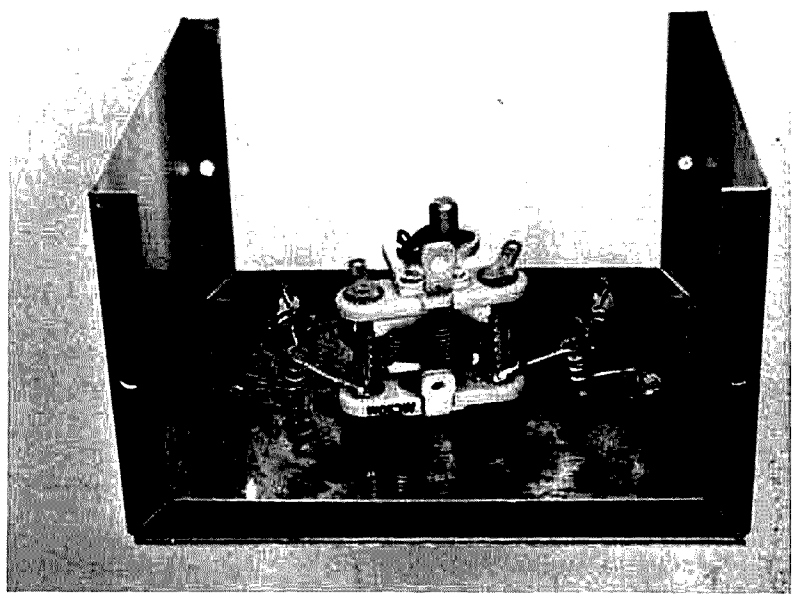


table 1. Bandpass filter data.

Filter	Frequency Range (MHz)	Frequency (MHz)	3 dB Bandwidth (MHz)	Loss
A	25 to 85	25	0.8	1.3
		32	0.9	1.1
		40	1.2	0.93
		60	2.0	0.80
B	52 to 190	55	6	0.40
		80	8	0.32
		110	10	0.30
		180	12	0.32
C	142 to 420	160	14	0.78
		210	20	0.70
		300	24	0.65
		380	30	0.56
D	230 to 1350	230	—	3.0
		250	—	2.5
		400	—	1.7
		432	45*	1.5
		480	—	1.4
		950	—	0.6
		1296	585*	0.35
		1350	—	0.35
E	200 to 2500	2300	400	1.0

* 20 dB bandwidth

Filter B tunes from 52 to 190 MHz.



bottom of the can (see fig. 5). Tuning is done with a threaded shaft and nut soldered in the center of the box; a friction lock nut is recommended. The last two filters don't have a calibrated dial, but once set they need be changed only slightly. A turns-counting dial with calibration chart could be used if desired.

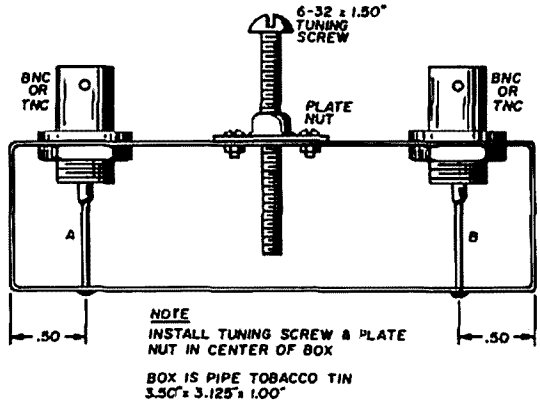
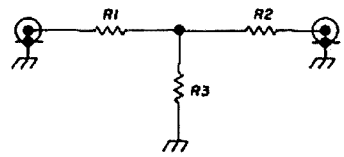


fig. 5. Bandpass filter that will tune from 2.0 to 2.5 GHz; filter will tune to 1296 MHz if a disc is attached to the end of the tuning screw. The position of the coupling probes A and B determines bandwidth and insertion loss. Space the probe 1/4" from end of tobacco tin for less bandwidth; however, insertion loss will be slightly increased.



Attenuation	R1, R2 (Ohms)	R3 (Ohms)
1	2.7	430
2	5.6	210
3	8.2	130
6	18	68
10	27	36
20	39	9.1
30	47	3.3

fig. 8. Attenuator circuit using standard 5% carbon resistors is not exact, but very close. Put all three half-watt resistors in a brass tube 2" long and solder BNC connectors to the ends.

application

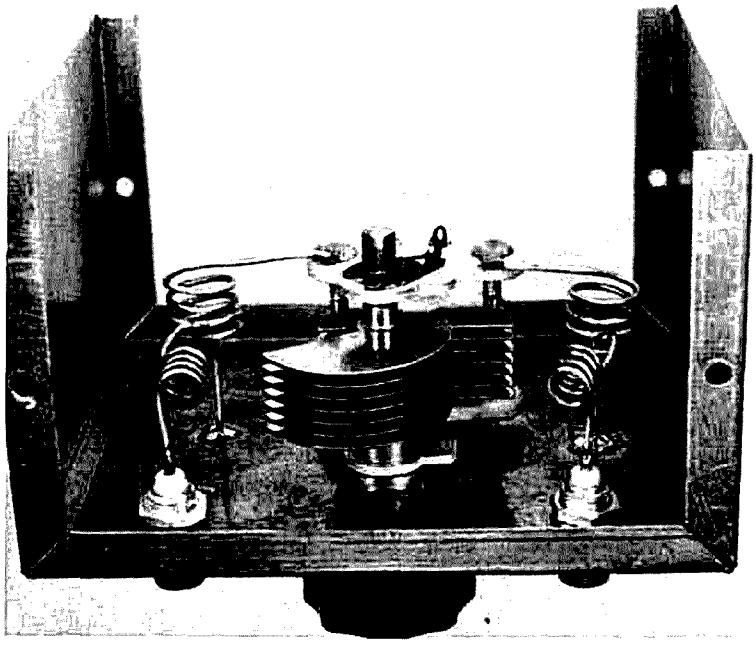
Using a filter at the output of a circuit will help you determine what signals are being generated. The filter will pass only the desired signal. Output level will then be the one desired; also by tuning the filter through the harmonic ranges, you can see what harmonic power is being contributed.

The output power of low-level amplifiers

fig. 6. Typical test setup using the bandpass filters.



Filter C tunes from 142 to 420 MHz.



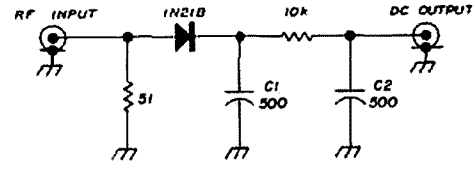


fig. 7. Sensitive rf detector probe. Capacitors C1 and C2 should be good quality ceramics.

or multipliers is sometimes only a few milliwatts, and of course most power-measuring devices won't indicate in this range. The only alternative is to use an expensive milliwatt meter or a home-made detector with a 50-ohm load. One is shown in **fig. 7.** I also recommend that an attenuator be used between the test circuit and the filter input as in **fig. 6** because the filter inductance can upset circuit tuning.

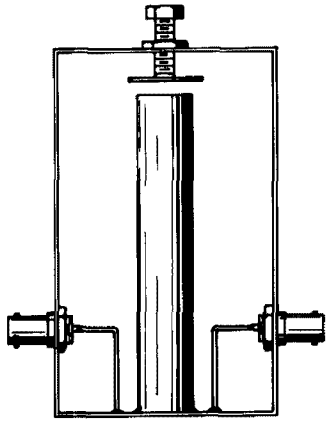
ham radio

single-pole bandpass filters

Fred Brown, W6HPH, Pine Cove, Idyllwild, California 92349

Probably the most common type of bandpass filter for amateur uhf work is the coaxial-cavity arrangement of **fig. 1.** This filter can be made to have very low insertion loss. However, the close proximity of the coupling loops results in some inductive coupling between input and output at all frequencies. The stop-band attenuation is consequently not as high as a single-pole filter can be.

fig. 1. The coaxial cavity filter commonly used at uhf.

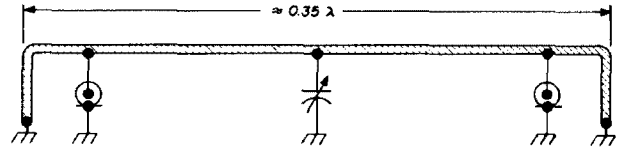


The arrangement of **fig. 2** places the input and output at opposite ends of the filter and results in better isolation. It also lends itself admirably to strip-line or trough-line construction. I built this filter to reject the usually strong, unwanted harmonics from a 432-MHz varactor tripler. Performance of the filter turned out to be quite remarkable, considering it was thrown together in twenty minutes total construction time with no attempt to optimize anything. Attenuation vs frequency between 200 and 1000 MHz is plotted in **fig. 3.**

construction

Construction details for the 432-MHz trough-line filter are shown in **fig. 4.** The trough is a 12-inch section of aluminum extrusion of the type used for sliding glass doors. The tuning capacitor is a miniature piston type with a range of 1.2 to 10 pF. It's a Triko 108-02M imported from West Germany. I was concerned that some of the 432-MHz insertion loss might be due to capacitor losses. To check this out I tried the filter in the output of the "big rig."

fig. 2. This trough-line filter is not only easier to construct than the cavity type of fig. 1 but also gives superior performance.



performance

At high-power level the lossy components in a filter will become hot. It wasn't possible to run more than 30 watts through the filter without the tuning capacitor arcing over, but at this power level there was very little capacitor heating. It did become just perceptibly warm to the touch, but so did the number 12

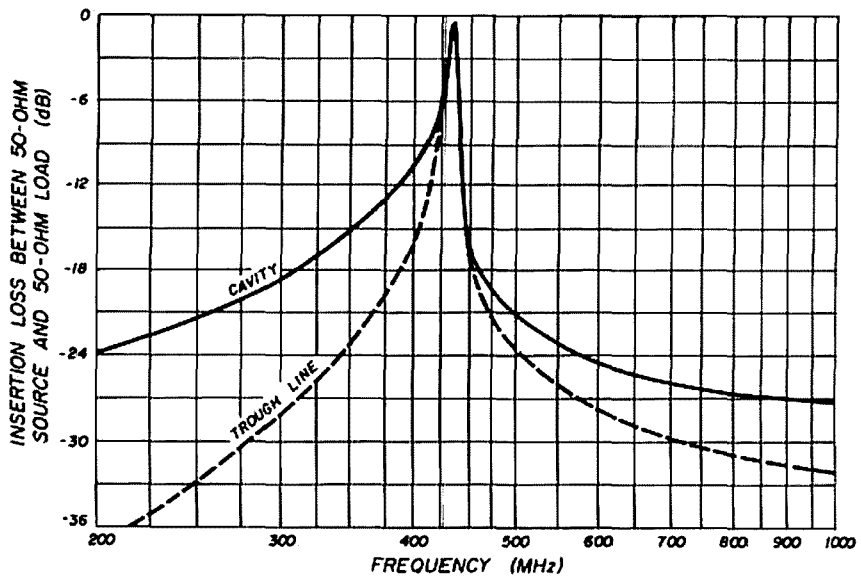


fig. 3. Comparison curves of the coaxial-cavity and trough-line 432-MHz filters. Note the superior stop-band rejection characteristics of the trough line.

copper wire center conductor. Probably the most important single factor for reducing insertion loss would be the use of a larger center conductor.

Also shown in **fig. 3** is the performance of a conventional cavity filter. It, too, was designed to reject unwanted harmonics from a varactor tripler. The filter was made by K6JYO from a surplus TS1/ARR-1 gold-plated cavity.

Note that the stop-band attenuation of the trough-line filter is better throughout the entire measurement range of 200 to 1000 MHz. The cavity filter did have slightly lower insertion loss: 0.4 dB as compared to 0.8 dB for the trough line. This is to be expected because of the higher-Q construction of the cavity. Loaded Q's of both filters were comparable, as indicated by the 3-dB bandwidth of 9 MHz for the trough line and 7 MHz for the cavity.

low-frequency version

The success of this uhf filter inspired construction of a low-frequency version using

The aluminum trough-line filter outperforms conventional coaxial cavity below.

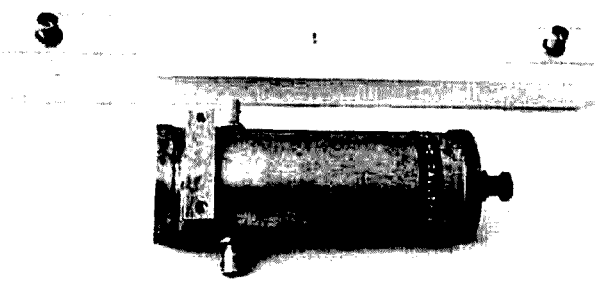
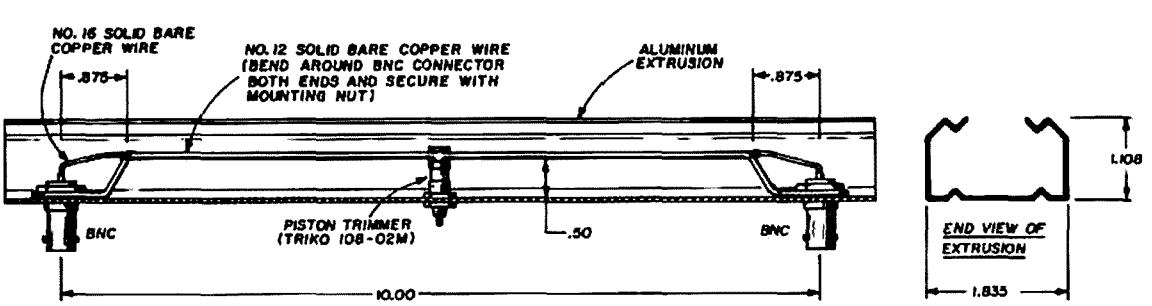


fig. 4. Construction details of the 432-MHz trough-line filter.



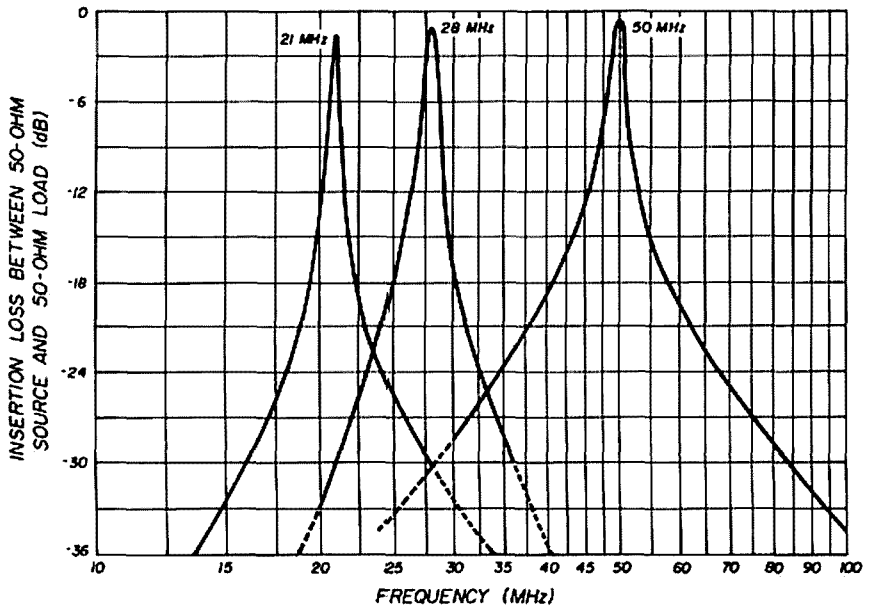
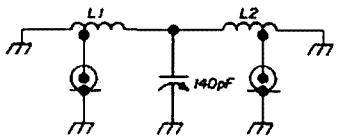


fig. 6. Performance curves of the low-frequency, lumped-constant version of the trough-line filter. Attenuation characteristics were measured with the filter tuned to three different frequencies: 21, 28 and 50 MHz.

lumped constants. It is built on a 2 x 4 x 8-inch chassis and consists merely of two coils, two BNC fittings, and one tuning capacitor. The schematic is shown in **fig. 5.** The LC combination gave a tuning range that covered the 15, 10 and 6 meter bands, so the filter response was measured at 21, 28 and 50 MHz.

fig. 5. Low-frequency version of the trough-line filter. L1 and L2 are 13 turns of number 16 baro copper wire, 5/8-inch diameter, 1-3/8-inch long, air wound, and tapped 1¼ turns from ground. The coils should be *oriented* to minimize inductive coupling.



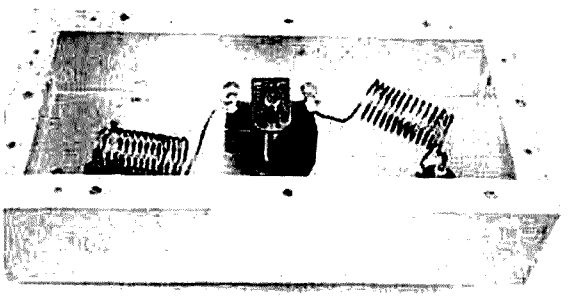
performance

As can be seen from the curves of **fig. 6,** this gadget should make an excellent tvi filter if you don't mind retuning whenever you change frequency. The 3-dB bandwidth measured 2.2 MHz at 6 meters, 600 kHz at 10 meters, and 400 kHz on 15. Insertion loss was 0.7 dB at 50 MHz, 0.8 dB at 29 MHz, and 1.2 dB at 21 MHz.

The insertion loss could have been reduced by using higher-Q coils (number 14 wire instead of number 16, for instance); or by tapping the input and output higher up on the coils, say two turns from ground. This would also increase the 3-dB bandwidth.

Power-handling capability will be determined by the voltage breakdown point of the tuning capacitor. I've run 40 watts into this filter at 29 MHz, resulting in 35 watts delivered to the load. At this power level there is no tendency for the small variable capacitor to break down, but the coils become warm to the touch, indicating this is where the other 5 watts are being lost.

ham radio



Low-frequoncy lumped-constant version of the trough-line filter. This filter can be tuned to 6, 10 or 15 meters.

standards for amateur microwave communications

This standard microwave system offers a practical means for amateur work above 1 GHz

Richard B. Kolbly, K6HIJ, 26334 Community Boulevard, Barstow, California 92311

In 1956, the San Bernardino Microwave Society was incorporated to further the art of amateur communications above 1000 MHz. A prime objective of this organization has been to develop equipment and techniques for amateur operation that can be duplicated by the average amateur experimenter at moderate cost. It soon became apparent to the members that it would be necessary to develop a set of standards for amateur microwave work if the art was to be developed in other than a haphazard manner.

This article, which I hope will be first of many in amateur microwave techniques, explains the working standards that the SBMS members have evolved for their own use. These have been used by the members for a number of years and have proved to be very practical in field operation. The standards pertain to operating frequencies, polarization, modulation, power supplies and component interconnections. The standard SBMS system is full duplex, allowing simultaneous transmitting and receiving, with the transmitter frequency removed 30 MHz from the receiver frequency.

The microwave amateur bands are very large compared with the lower bands. In order to communicate effectively over long distances, it is necessary to have certain standard frequencies and operating techniques that are understood and used by the station on the other end. For example, the 3300-MHz band (high S-Band) is two-hundred megahertz wide—far too wide to search for a weak signal with a 1-kHz bandwidth receiver!

A large amount of surplus microwave equipment is available on the market, which is suitable for amateur work, with a wide range of possible interconnections. The members of the SBMS felt it was necessary to standardize interconnections to allow rapid replacement of system components.

a typical station

The SBMS members operate a full-duplex system, with the transmitter also acting as the receiver local oscillator. The transmitter is frequency-locked to a crystal-controlled reference oscillator for stability. **Fig. 1** is a block diagram of a typical SMBS microwave station. The antenna feed also acts as a receiver mixer. This system is designated ROCLOC, meaning **Relative Or Crystal Local Oscillator Control.** It was developed primarily by Don Thompson, W6IFE, and George Tillitson, K6MBL. This ROCLOC system has proved to be an effective and versatile ham microwave setup. Interconnections are more or less standardized to allow equipment to be easily interchanged and tested.

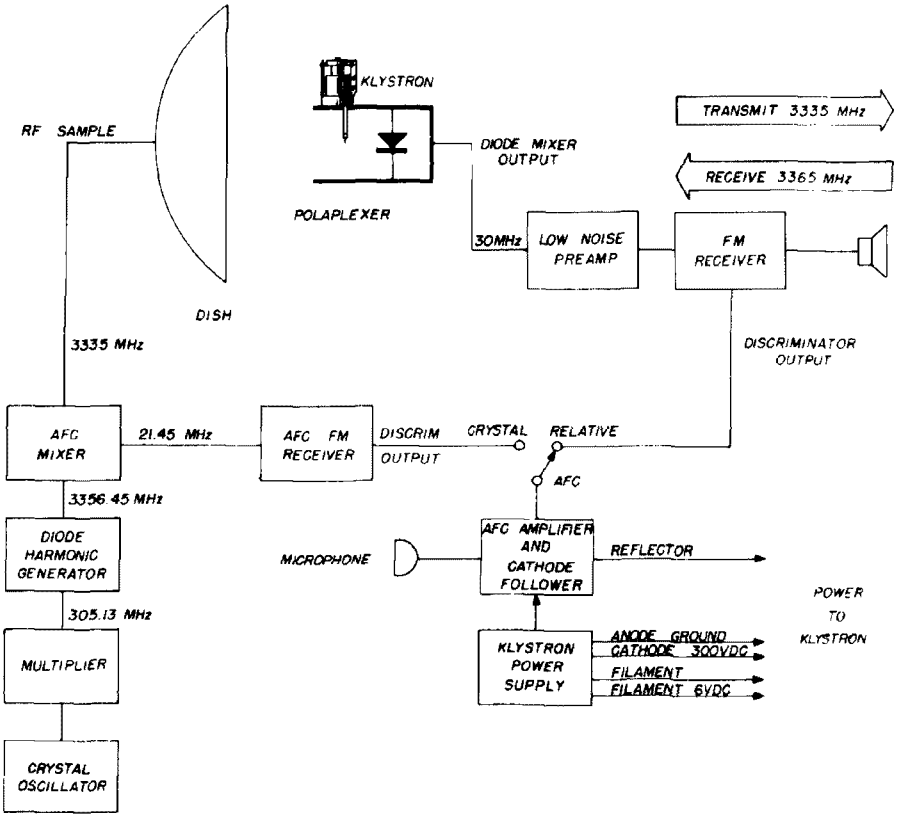


fig. 1 A 3300-MHz ROCLOC station as used by members of the San Bernadino Microwave Society, Inc. (from W6OYJ).

frequencies

Standard operating frequencies are a necessity, as most members operate with narrowband receivers that have a limited tuning range. The selection of standard frequencies for microwave hamming was based on readily available klystrons and a safe separation from the band edges.

The selection of the frequencies listed in **table 1** was based on a 5-MHz safety factor to prevent out-of-band operation of a station accidentally tuned to the other side of an incoming signal. For example, W6IFE transmits to K6HIJ on 3335 MHz. K6HIJ is supposed to have his transmitter on 3365 MHz, 30 MHz away, but since his receiver has no image rejection, he accidentally tunes his transmitter to 30 MHz **below** W6IFE. He is on 3305 MHz, still 5 MHz inside the amateur band. All SBMS work is based on an i-f of 30 MHz.

Since the transmitter polarization is at right angles to the receiver polarization, (standard polaplexer), all SBMS members polarize transmitted and received signals 45 degrees from vertical, with the transmitted signal polarized 45° to the right of vertical in the direction of propagation.

klystron voltages

A large number of klystrons available on the surplus market will operate satisfactorily with 300 volts of beam voltage, so the SBMS has standardized on this beam voltage, with the positive terminal (klystron body) at ground potential. Also, direct-current heater operation has been found necessary. Therefore a low-ripple, dc filament supply (5.8-6.3 V; 1.5 A), isolated from ground is required. The klystron reflector voltage (referred to the klystron cathode) will operate from −100 to −300 volts at very low current drain. Ripple, hash and noise on all power supplies should be less than ten millivolts (0.01 volt). The reflector supply is most critical in this respect.

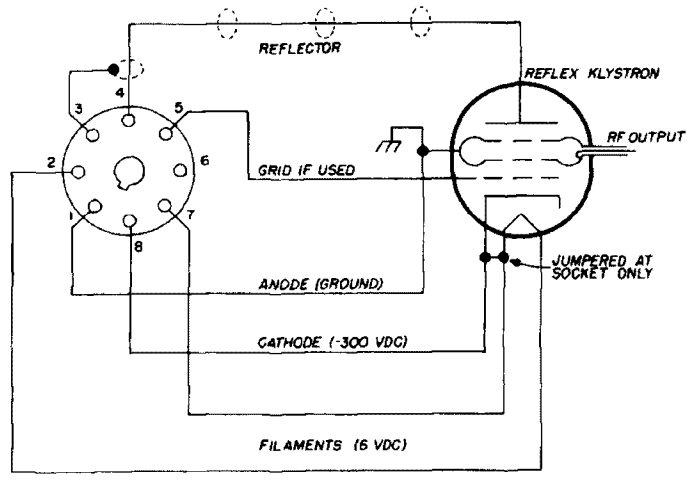


fig. 2. Typical klystron power connections.

For power connections to the klystron, standard octal plugs and sockets are used. The pin connections are shown in **fig. 2.** The supply side of the connection should be the female connector. By using these socket connections, it's possible to interchange klystrons with a minimum danger of damage to the klystron or power supply or of improper klystron operation.

modulation modes

Amplitude moduation of a reflex klystron is difficult, so frequency modulation is used by the SBMS. For wideband operation, deviation is set to be compatible with the receiver, usually ±50 kHz, as most of the members use BC-683 fm tank receivers as 30-MHz i-f strips. For weak signal work, it's necessary to go to narrow bandwidths and smaller deviation.

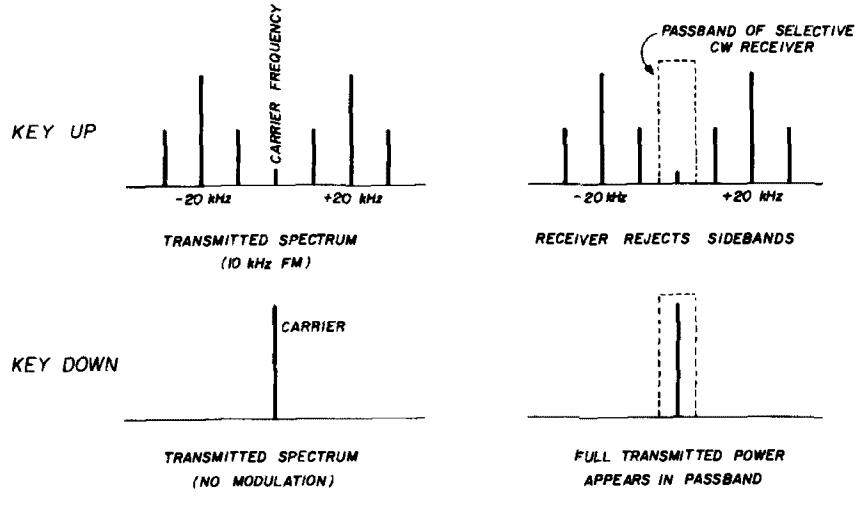


fig. 3. Principle of pseudo-cw modulation.

table 1. Standard amateur preferred microwave frequencies.

band	band limits (MHz)	preferred frequencies (MHz)
low S	2300-2450	2385 & 2415
high S	3300-3500	3335 & 3365
C	5650-5925	5860 & 5890
X	10,000-10,500	10,035 & 10,065
K	21,000-22,000	21,035 & 21,065
		21,935 & 21,965

Usually the receiver has a bandwidth of 1 to 2 kHz, and deviation is set accordingly.

For very long-haul, weak-signal work, cw offers the best performance. To accomplish this form of modulation, the SBMS has developed "**pseudo-cw**" modulation. **Fig. 3** illustrates the principles involved. The transmitter is frequency modulated at a high audio rate (10 kHz or so), and the deviation is adjusted to the first carrier null, with all the transmitter power in the sidebands. A narrowband receiver tuned to the carrier frequency will not receive a signal, as the sidebands are outside the receiver passband.

This modulation corresponds to a "key up" condition. Depressing the key ("key down") removes the modulation from the signal, and all transmitter power appears on the carrier frequency. This "**pseudo-cw**" signal is received exactly as a standard cw signal, with the receiver bfo on.

This technique has been as effective as standard "on-off" cw modulation for long distance work yet allows the klystron to be frequency stabilized at all times.

conclusion

I have attempted to outline the standards for amateur microwave work used by the members of the San Bernardino Microwave Society. Using these techniques, communications have been accomplished over long, indirect paths over mountainous terrain. We've observed that a frequency change of one megahertz will result in a 20-dB change in signal strength over an indirect path. The **ROCLOC** system allows rapid changes in transmitter frequency with stable operation. (Any amateur system should have the capability of making frequency changes.)

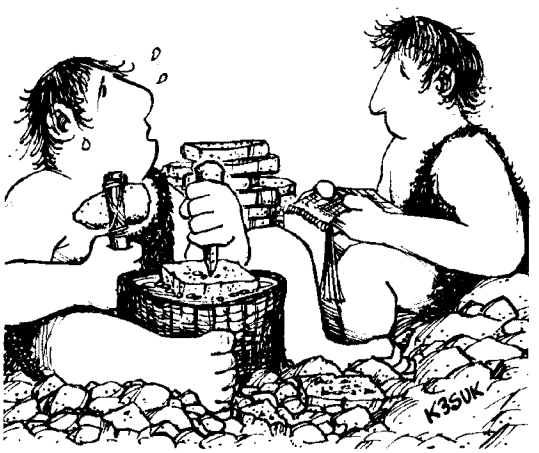
In review, the standards used by the SBMS are:

operating frequencies	refer to table 1
intermediate frequency	30.0 MHz
polarization	linear, 45 degrees, transmit and receive
beam voltage	300 volts, ±1%, less than 10 mV ripple
fm, wide	±50 kHz
fm, narrow	±1 kHz
morse code	"pseudo cw" (**fig. 3**)
rf-tuning range	greater than ±1 MHz

If you are planning to move up to the microwave bands, I'm sure you will find these operating standards practical.

ham radio

a look at those days before spark... b.s.



"I love to operate . . . but these QSL's are pure drudgery."

solid-state modification of a mobile converter

An easy way to modernize a Gonset tube-type converter for mobile use

John R. Schuler, P. O. Box 37, Hanover, Maryland 21076

I was thinking of installing an old Gonset converter in my car, but the car's 12-volt system presented some problems. The converter tubes had 6-volt filaments, which would require a dropping resistor or heavy-duty zener. Also a power supply of +180 volts or more would have been necessary. Both these items would have represented considerable wasted power, more expense, installation problems, much more work, noise pickup problems, hash filtering, etc. Besides, I had a new all solid-state bc radio.

I felt that fet's and/or transistors would eliminate all these problems, and the small amount of filtered dc power required could be readily supplied by the bc set after it had been adequately filtered to eliminate input noise. Additional benefits would be instant warmup, no oscillator drift, no tube-pin noise from road shocks, and no worry over failure of tube filaments, element shorts, etc.

If mechanical redesign could be eliminated and electrical redesign limited to changing a few resistors and capacitors, the overall expense and effort would be drastically reduced even further. With this in mind, a most satisfactory modernization was achieved with equal or better performance than when the converter left the factory.

construction

A most troublesome problem after removal of the tubes was finding a place to mount the transistors, since clearance was negligible between the tube socket connection ears and the bottom of the enclosing chassis. I finally decided to "spread-eagle" the short wires of the solid-state devices, use 1/4-inch pieces of spaghetti, then solder the wires directly to the socket ears where old and new components would also be convenient for connection. An all-fet converter was originally planned. However, the great improvement in conversion gain afforded by an npn transistor justified its use as the mixer.

The modifications are shown in **fig. 1.** I found it necessary to put in a slide switch to disconnect the antenna circuit from the output fet (during bc reception only), otherwise the loading effect was sufficient to ruin the Q and gain of the bc set. But this was the only mechanical change. There was plenty of room on the back apron of the converter to install this switch.

Where values of components are not given on the schematic, the value is the same as originally furnished by Gonset. Of course, the filament wiring, screen-grid wiring, screen-grid bypass capacitors anl dropping resistors may be cut out of the chassis; and good riddance. Note that the pilot lights were changed from 6-volt 47's to 12-volt types and disconnected from the line feeding the collector and drains. They may be separately fed from a battery clip and wire attached to the instrument panel light to lessen the IR drop from the bc set hash filter. In this way, when the bc set is turned off, the converter is "fail-safe" turned off too, irrespective of converter switch position. Wire from the on-off switch on the bc set should be shielded and fed to the shielded B+ line of the converter via phono pin and jack, or similar means, so each may be separated at will.

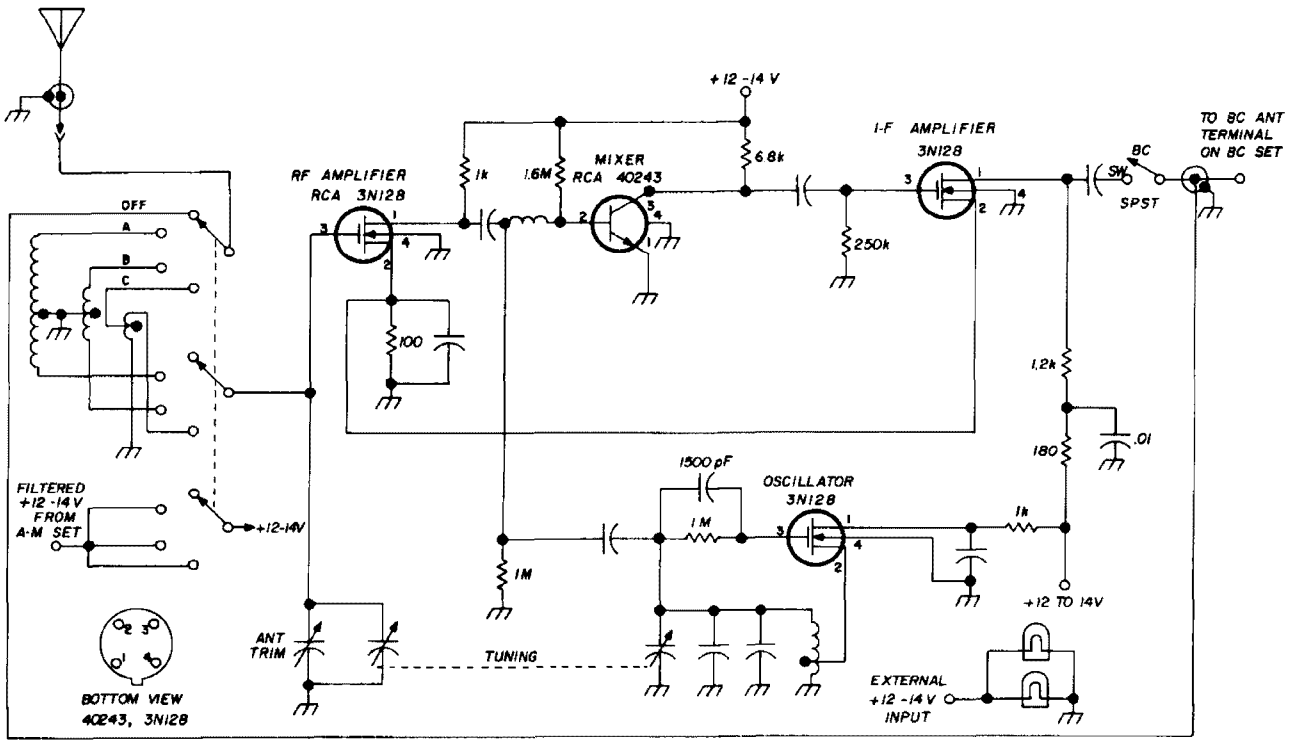


fig. 1. The solid-state Gonset converter. Mosfet's and an npn transistor eliminate tube problems of noise, drift and high power drain. Components without values are unchanged from the original converter.

affect of mismatched transmitter loads

Does the character of the load affect power amplifier efficiency?

Carl Drumeller, W5JJ, 5824 N.W. 58th Street, Warr Acres, Oklahoma 73122

Assume, if you will, that a radio transmitter feeds a "lossless" transmission line terminated in a non-reactive load equal to the characteristic impedance of the line; the vswr will be unity, of course. Further assume that the transmitter runs 1000 watts input power and operates with an efficiency of 60%. A through-line wattmeter in the transmission line would then indicate 600 watts power output. For a final assumption, let's say that the plate of the vacuum tube in the final rf amplifier stage shows a faint trace of "color" under these operating conditions.

Now remove the non-reactive load and replace it with a complex load—partially-resistive and partially-reactive. Also, its total impedance is not equal to the characteristic impedance of the transmission line. This, of course, upsets the tuning of the transmitter; so quickly retune it to resonance and reload the final to draw 1000 watts dc input power.

the question

Now the question: will the power amplifier plate show less, the same or more "color?"

This question will educe a lot of discussion among technical people—even a bit of argumentation. Opinion, even among well-qualified engineers, will be sharply divided. Most agree that a final solution can be reached only by setting up an experimental circuit and observing the results under carefully controlled conditions.

the conditions

Pending such an empirical solution, however, there are several thought-provoking aspects that should be considered. First, let's

tie down some conditions. Assume that the unloaded Q of the plate tank circuit is very high; that the tank circuit normally has a loaded Q of 12; that the inductor of the tank has low resistive losses; that a moderate change in the inductance of the tank inductor will produce no change in its total resistance; that the transmitter is operating on a relatively low frequency in the high-frequency range so that tank losses are mostly concentrated in the inductive leg of the tank; that the variable tuning element in the tank circuit is a variable inductor of the variometer type. Having assumed all this, we're ready to look at the schematic diagram in **fig. 1.**

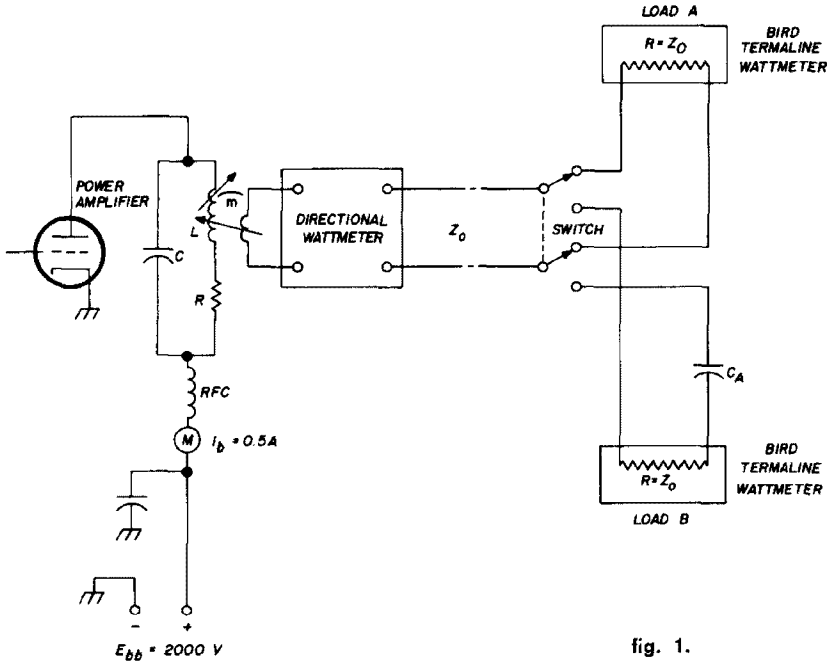


fig. 1.

You will recall that the question stipulated an over-all efficiency of 60% with a purely resistive load. The important factors determining the degree of operating efficiency are grid bias, rf excitation and the plate load. We shall assume that grid bias and rf excitation have been adjusted to optimum and will remain unchanged throughout the test. The third factor, plate load, is a variable depending upon the loaded Q of the plate tank circuit. Assuming the capacitor has negligible contributory loss, we can say that tank Q is equal to the inductor's reactance divided by its resistance ($Q = X/R$). This Q, in turn, determines the impedance into which the tube works. When the tank circuit is tuned to resonance it may be stated that

$$Z = \frac{X^2}{R} = XQ$$

Keep in mind that the capacitance of the tank circuit is fixed. Therefore, for a state of resonance ($X_L = X_C$) at a given frequency, the effective value of inductance must be constant. This is important, for we shall be tuning the variometer to maintain a status of resonance. Now slip back a notch and consider the matter of impedance. Look at the formula $Z = X^2/R$—if X_C is fixed and X_L must always equal X_C, it's evident that resistance is the only variable that can be manipulated to attain a desired value of impedance. We shall assume we have reached that desired value when the dc plate current has a value of 0.5 ampere with the tank circuit in resonance.

Since we stipulated that the unloaded Q of the tank circuit is high (same as saying that the inductor's resistance is low), the effective resistance of the tank must actually be predominantly made up of resistance reflected from the load through the mutual

coupling, m, that links the tank inductor to the transmission line pick-up coil.

the experiment

Connect load A to the transmission line, fire up the transmitter, and adjust the plate variometer and the mutual coupling, m, so that when in resonance the dc plate current is 0.5 ampere. The directional wattmeter indicates 600 watts incident power and zero reflected; the meter on the Termaline® wattmeter used for load A also indicates 600 watts. The extent of the glow on the plate of the final amplifier tube is carefully observed and recorded. This concludes the first step.

For the second step, the transmission line is shifted from load A to load B. Load B consists of a Termaline® wattmeter in series with a capacitor. The total impedance of this load is expressed by:

$$Z = \sqrt{X_C^2 + R^2}.$$

It is evident that the new load is not matched to the transmission line so there will be standing waves present on the line. To simplify matters, let's assume that the transmission line is a half-wave long (at the operating frequency). Then the capacitive reactance (X_C) of the load will be present at the sending end of the line. It will be reflected into the inductive leg of the plate tank circuit as a value of X_L, causing the total X_L to be too great to maintain an equality with the fixed value of X_C. Therefore, the plate tank will not be in resonance. You quickly restore it to resonance by adjusting the variometer. Now total tank-circuit X_L is back to its original value.

Although the value of resistance in the load has not been changed, it is probable that m will have to be altered to maintain the tank R at a value that will permit a dc plate current of 0.5 ampere at resonance; make the necessary adjustment.

the results

The tube now has a dc input power of 1000 watts. The directional wattmeter in the transmission line will now indicate something other than 600 watts incident power; also, it will show some reflected power. The question, though, is what has happened to plate dissipation? The answer is short and sweet: nothing!

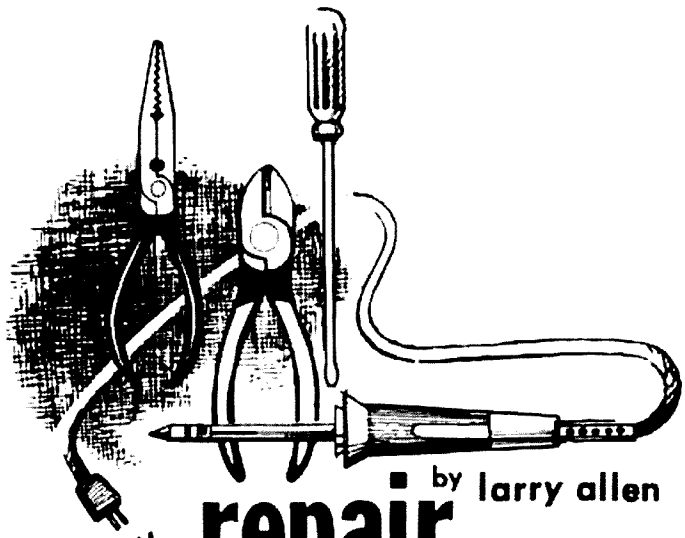
This is because the power generator, the tube, sees exactly the same load it saw when the transmission line was terminated with a resistive load equal to the characteristic impedance of the transmission line. This condition was established when the plate-tank variable inductor was returned to resonance and the coupling adjusted for the same dc plate current.

The plate tank circuit is the key to the situation. It performs three functions: it provides an optimum value of non-reactive load for the generator to work into, it provides frequency selection by offering this load only at the selected frequency, and it serves as a transforming device, transliterating random loads into the load that is optimum for the power generator. It is the third function that seems to be least understood.

Although a rather unusual type of tank circuit has been used in this illustration, other tank configurations work equally well and perform precisely the same service. The sending-end impedance can be (and usually is) almost anything but the 52 ohms resistive that the transmitter is designed to work into. The impedance may vary over the range from 15 to several hundred ohms—and it may be purely resistive, resistive plus inductive or resistive minus capacitive! This is what the tank circuit must cope with to perform its vital third function.

Fortunately, almost any properly designed tank circuit is capable of making this transformation. The limitations are mostly those that are built in by the designer's choice of components. There are two factors that have a lot to do with this choice: cost and ease of tuneup. The first is a matter of keeping the selling price low to compete in a highly competitive market. The other is to keep tuning procedures simple so the equipment won't be damaged during tuneup. If the tubes go out in a blinding flash or the tank circuit melts down into a sodden mess, either the designer cut too many corners or the operator lacked the skill to make the necessary corrective adjustments. Consider that before you cuss the designer!

ham radio



by larry allen

the repair bench

putting a scope to work in ham gear

Here I am to continue with scopes. I told you last month how to fire up the scope and get displays locked into position on the screen. This month I'll give you some specific uses for it on the ham repair bench.

A scope, remember, lets you **see** the exact shape of an ac waveform. You can identify sine waves, square waves, tv sync pulses or any unusual waveshape. More than that, and often more important, you can measure peak-to-peak values of waveform voltages, whether they're nice round sine waves or crazy, misshapen oddballs.

measurement methods vary

Probably the chief difference in various brands of service-type scopes lies in how you measure voltages with them. Yet, they're all fairly easy. There are three main ways. Each uses the scope's variable **vertical input** control, but in different ways. The **vertical input** switch is always a decade-type multiplier. And the marked graticule on the face of the scope is important to voltage measuring.

In one method, you first feed in a signal of known value and adjust the **vertical input** control to make the trace coincide with that voltage marking on the scope graticule. That calibrates the input amplifier. You can thereafter read the voltage value more or less directly on the scope graticule.

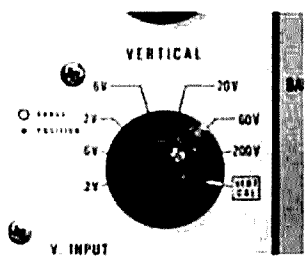
The second way of measuring peak-to-peak voltage is almost identical, except that the calibrating voltage is inside the scope. You just turn a switch to **calibrate,** and adjust the **vertical input** knob till the waveform display fits between two "calibrate" lines on the graticule. Then, you can read any signal-voltage value from where it fits on the graticule. The switch and the waveform are shown in **fig. 1.**

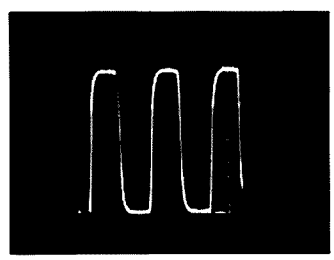
You've probably figured out that you can't mess with the **vertical input** knob once it's set. If the display is too small, you turn the input multiplier switch to a more sensitive position. If the display is too large (off-screen), switch to a less sensitive multiplier.

The third common method of measurement lets you read the voltage value directly from the **vertical input** knobs. **Fig. 2** illustrates the controls. You adjust the **vertical input** controls to make the waveform display exactly 2 inches high on the graticule (which is marked in inches). Then, you read the number pointed to by the variable knob and multiply it by the multiplier setting. That tells you the peak-to-peak voltage of the input waveform, no matter what its shape.

The waveforms that follow are photographed from a B&K model 1450 servicing scope. Its lighted graticule, shown in **fig. 3,** has two numbered scales: 0-2 and 0-6. The multiplier switch determines which scale

fig. 1. Before measurements, the vertical input amplifier must be calibrated to a standard voltage. Switching to vertical applies internal signal to deflection amplifier in this scope. Display is adjusted to fit the full-scale graticule mark.





lights up. The correct full-scale voltage is listed on each waveform photo so you can read peak-to-peak amplitude accurately. As you go along, reading voltages on a scope like this becomes second nature.

tracing input power

A simple use for the scope, and a good one to practice with, is checking input ac voltage to a receiver or transmitter. It's the same thing you'd ordinarily do with your ac voltmeter.

The diagram in **fig. 4** shows the test points in a popular ssb transmitter. The photo below the input wiring shows the waveshape. Ac line voltage is a sine wave.

The 120 volts your voltmeter reads is a root-mean-square (rms) value, and the scope shows peak-to-peak (p-p) values. Line voltage is 335 volts peak-to-peak. The highest full-scale voltage reading on this scope is 200 volts, but its probe has a 10X switch on it. That attenuates any signal voltage by a factor of 10.

The multiplier switch for the photo in **fig. 4** is set at 60. A voltage of 600 will therefore reach all the way to the top of the 6 scale. The waveform you see goes about a subdivision and a half above the 3 mark.

The **sync** is set for **—int,** but it could just as easily be set for **line;** either would lock the line-voltage waveform in tightly.

The **sweep** control must be set for 10–50 Hz, and the **fine frequency** adjustment is turned so four cycles appear on the display. Adjust positioning controls so the bottom of the waveform is on the base line of the graticule.

Use an isolation transformer between the plug of any chassis you're testing and the power line. Otherwise, you can blow fuses with a direct short between the scope ground lead and one side of the power line. (Not to mention the big chance of a nasty—even deadly—shock.)

Clip the ground lead of the scope to the ground side of the power cord. Then, move the test probe from point 1 to point 3 as shown. You should get the same waveform at every point. If you get a waveform when you probe the grounded end of the transformer primary, however, it's a sign the ground is open. You can test the primary winding, too, by opening its ground connection and probing at the ground end; if you get a waveform, the primary is okay (**fig. 4** shows this test).

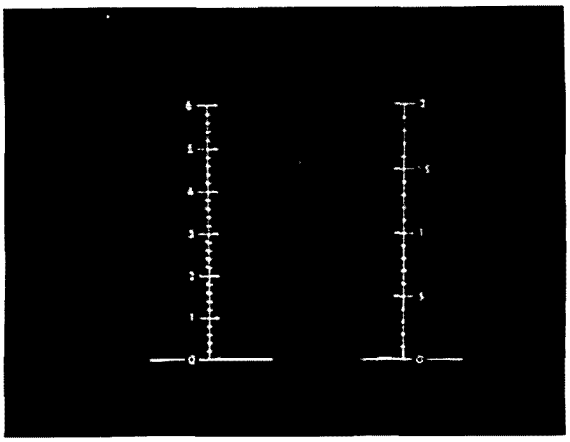
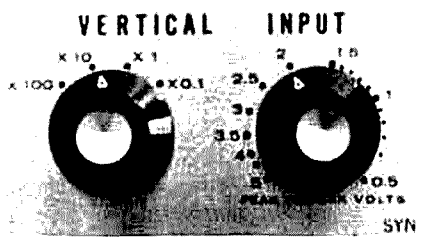


fig. 3. You read p-p voltages on scales with this scope, much like you do with a voltmeter. Either scale is used with decimal multiples chosen by the *vertical input* switch. Scales light one at a time.

fig. 2. Waveform peak-to-peak amplitudes can be read directly from the knobs of this scope. The display is made exactly two inches high on the graticule by turning the knobs; then voltage is read.



You can test the secondaries, too. Just probe points 4, 7, 9, 12, and 13. Remember that the **Vac** readings listed on a schematic are rms values. To know what p-p reading you should expect, multiply rms voltage by 2.8. Probing at point 10 also tests the doubler filter that couples ac voltage to the first rectifier.

power-supply filter capacitors

A scope is a fine tool for testing power-supply filters. It lets you actually see how much ripple is left in the dc output of the supply.

Here's a good rule-of-thumb: Ripple should

not exceed 1% of the dc voltage across the filter, **with the equipment drawing full load.** Thus, in a 500-volt dc supply, output ripple under load should be no more than 5 volts. If it is, the filter capacitors aren't doing their job.

For a regulated supply, the ripple percentage should be even lower. The regulator circuit or its capacitors are faulty if the ripple at the output exceeds 0.1% of the dc value. Ripple voltages are always measured peak-to-peak.

The waveform photo in **fig. 5** is taken across the filter of a 400-volt dc power supply. The full scale is 20 volts, so the ripple waveform is 4 volts p-p. The scope **sweep** is set for 10–50, and **fine frequency** is set to display four cycles. **Sync** is **–int.**

sine waves in amplifiers

Here is where a scope begins to shine. It can show up distortion at the same time it's measuring gain in an amplifier. Suppose you suspect trouble in the speech amplifier section of a transmitter.

You feed about 50 mV of audio signal into the microphone input jack (**fig. 6**). Any frequency around 1000 Hz works fine. Then you move the scope probe from stage to stage, at inputs and outputs, all the way to the modulator. Each amplifier stage should multiply the signal voltage by at least 10 or 20.

A clipper, on the other hand, is supposed to prevent overmodulation. It limits how much audio signal reaches the modulator. You can check the clipper with a scope.

Connect the probe at the modulator grid or at the output of the clipper stage itself. Turn up the audio generator output, watching the scope as you do. When the audio signal gets strong enough to cause 100% modulation, the clipper starts acting. The sine wave begins flattening out as the second photo of **fig. 6** shows. From that point on, you should be able to turn the generator wide open without getting a higher waveform on the scope. The shape only gets flatter.

You can trace rf amplifiers with the scope, too. Just feed in a modulated rf signal, and use a demodulator probe with your scope.

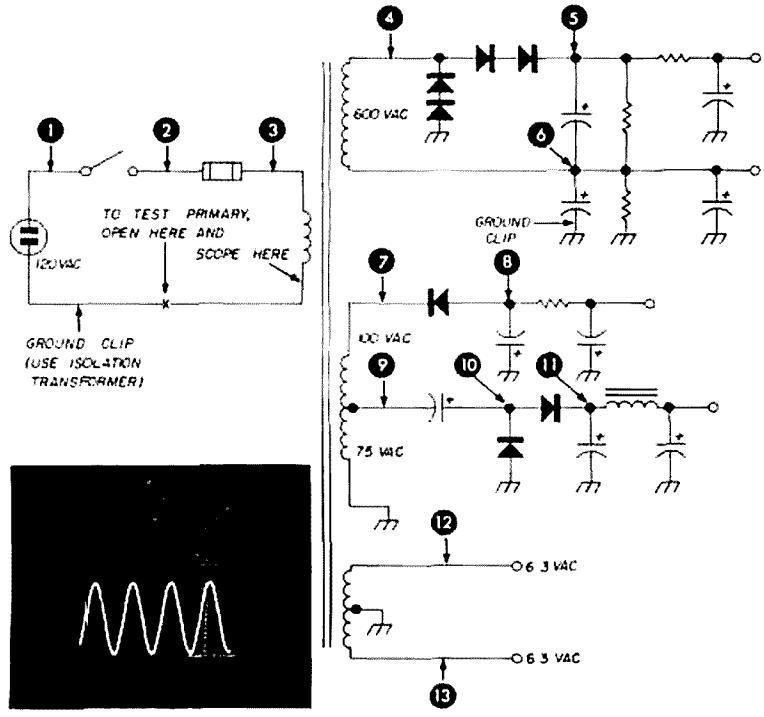
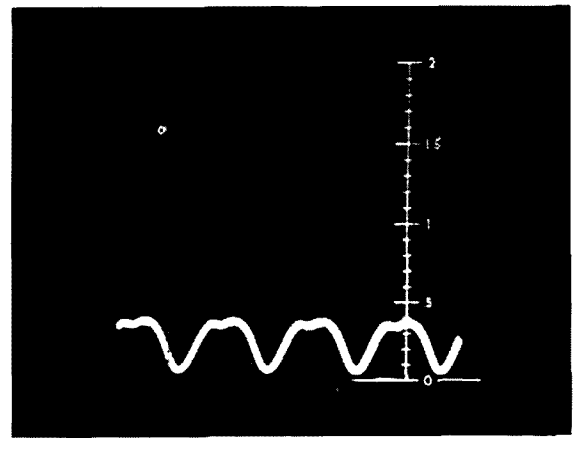


fig. 4. Power-supply test points. Waveform of input voltage. Full scale in this photo is 600 volts; peak-to-peak voltage therefore appears to be about 335 volts.

With amplitude-modulated signals, any circuit fault that cuts down the rf signal also cuts down the audio modulated on it.

You can even peak up receiver alignment with a scope fed by a demodulator probe. Just use the amplitude of the waveform on the scope as an indicator of peaking as you turn the adjustments. You can move the

fig. 5. Ripple voltage found in one lullwave power supply. Taken across input filter. Full scale of graticule is 20 volts; 4-volt p-p reading indicates filter capacitor is doing its job fairly well.



probe from stage to stage as you progress with tuning adjustments, or you can connect it following the last rf or i-f stage. If you go beyond an a-m detector, you don't need the demodulator probe. However, you can align the i-f and rf stages in an fm or ssb receiver using an a-m signal and the demodulator probe with a scope.

ham television

In recent years, hams have found it logical to stick with the commercial tv standards. Thus, they can use standard receivers with modified tuners. Also, surplus and used transmitting equipment is more available.

For examining and measuring signal voltages in television stages, a scope is indispensable. The photos in **fig. 7** are the most common waveforms.

Figs. 7A and **7B** are photos of the signal voltage that contains both the video that makes the picture and the sync pulses that keep it steady. The scope probe is connected to the output of the tv set's video detector. Full scale of the graticule when these photos were taken was 6 volts p-p; the waveform is just under 5 volts p-p.

In **7A,** the scope sweep is set for about 5 kHz; that shows three cycles of the signal at the 15,750-Hz line rate. **Fig. 7B** is the same signal, but with the scope sweep set for about 20 Hz. That displays three cycles at the 60-Hz frame rate. Peak-to-peak voltage is the same, no matter what display rate is used.

From the vertical sweep section of the tv set, the signal voltage in **fig. 7C** is taken. It is a trapezoid, displayed on the 600-volt scale. Its amplitude is 220 volts p-p. The scope sweep is set for 20 Hz.

Fig. 7D is also a trapezoid, this time taken in the horizontal sweep section of a television receiver. The scope is set to sweep at about 5 kHz, to display three cycles. The full-scale voltage range on the graticule is 200; the waveform is 120 volts p-p.

You'll discover that waveforms in any television receiver are viewed at either of just two scope-sweep rates. The 5-kHz sweep rate is for all waveforms that bear any relation to the horizontal line rate of the tv receiver. The 20-Hz sweep rate is for waveforms that relate to the vertical frame rate.

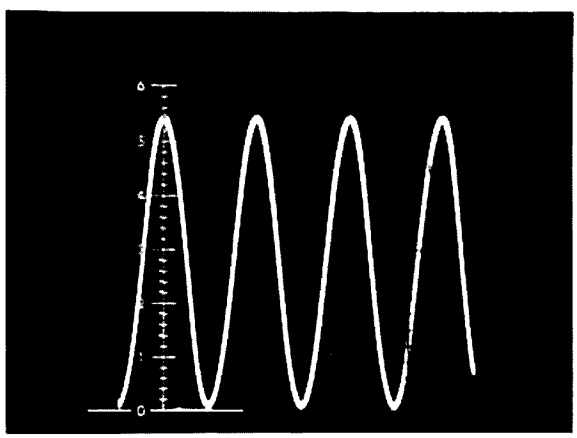
Fig. 7E is a special waveform found in the automatic frequency control circuit of the tv-set horizontal oscillator. The scope sweeps at the line rate. The graticule is at 200 volts full scale; the waveform is therefore 100 volts p-p in amplitude.

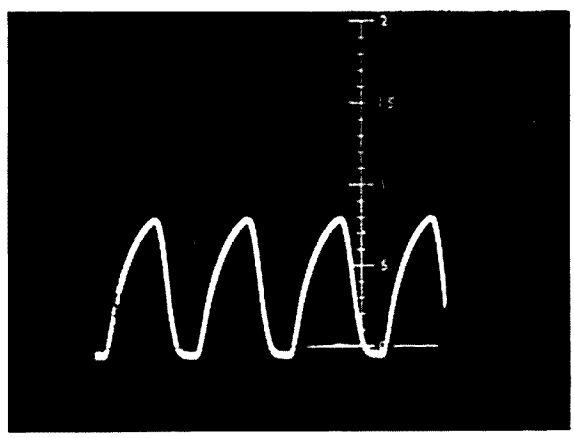
Another special circuit in TV receivers is the automatic gain control. In most modern sets, it is a keyed type. That is, a pulse is applied to it from the horizontal sweep transformer (called the flyback). The photo in **fig. 7F** is a sample of the keying signal voltage. The scope is set to display horizontal-rate signals. Peak-to-peak amplitude of the keying signal is 175 volts.

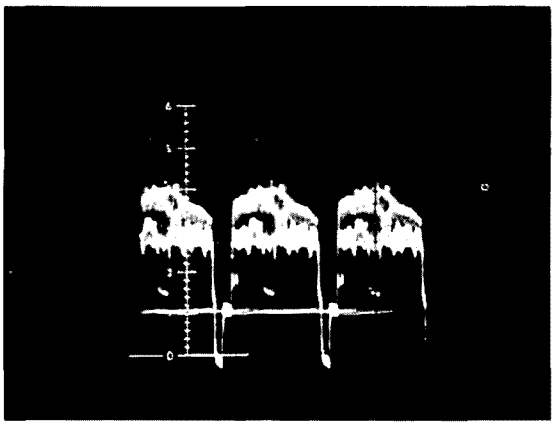
next month

These last two installments have given you a sample of how you can use a scope on the ham repair bench. A little practice makes

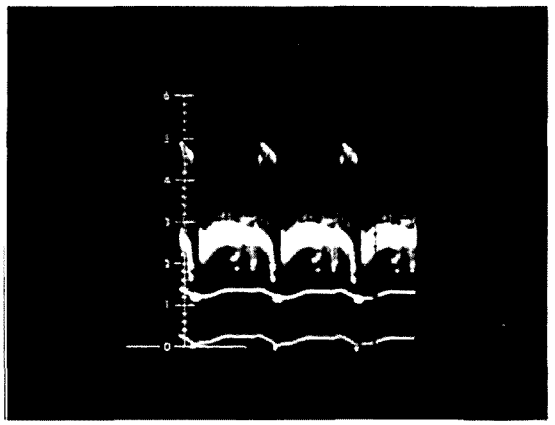
fig. 6. Sine-wave audio signal gets flattened in clipper when too much signal is fed in. That's normal, because clipper should keep signal at modulator from exceeding voltage for 100 percent modulation.



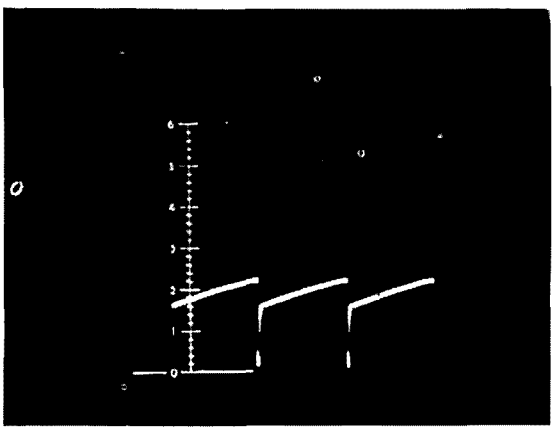




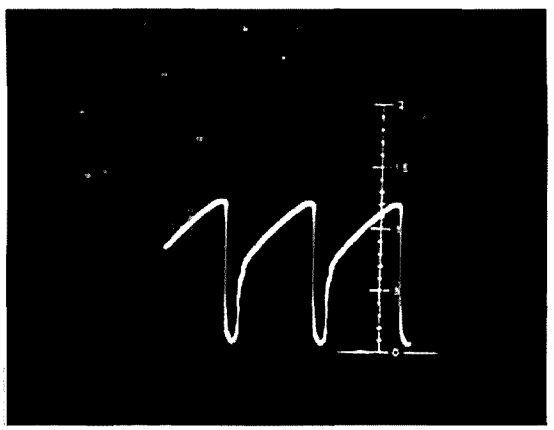
A. video waveform at line rate



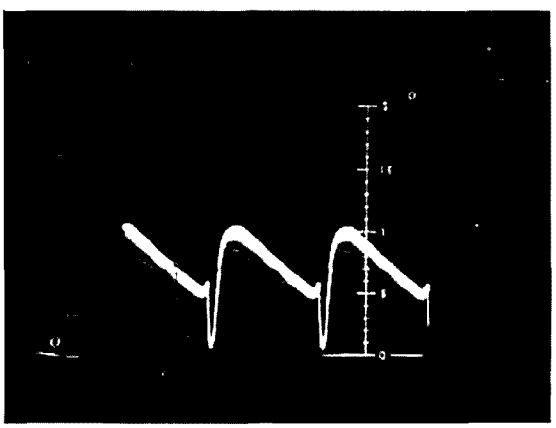
B. video waveform at frame rate



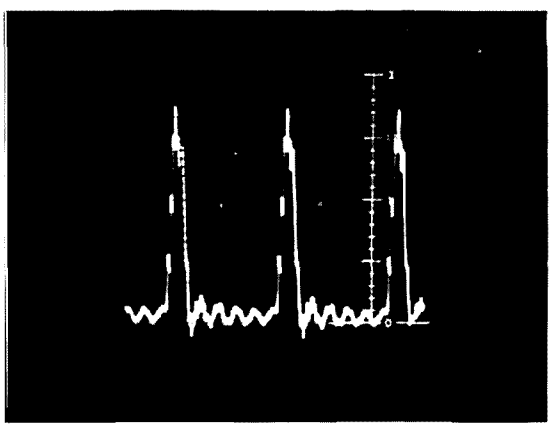
C. vertical oscillator waveform



D. horizontal drive signal



E. afc comparison signals



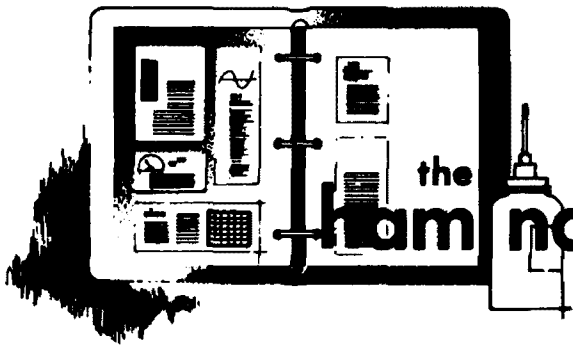
F. agc keying pulse

fig. 7. Typical waveforms from television receiver that can be used for amateur tv reception. Video waveform, displayed at line rate (A), video waveform, displayed at frame rate (B), vertical oscillator waveform, at frame rate (C), drive signal to horizontal output tube, displayed at line rate (D), waveform of comparison signals in afc stage, shown at line rate (E) and keying pulse for agc stage, at horizontal rate (F).

you as familiar with your scope as you are with your voltmeter. And you'll find the scope is far more versatile.

Next issue, I shift to a new topic: single-sideband. Tuning up an ssb transmitter has some things in common with any other ham transmitter. But, there are important differences. Those are the subject of **repair bench** next month.

ham radio



mono-loop antenna

Nothing new or startling is claimed for this antenna. However, like the open-wire-line fed flat top, the one-wavelength loop deserves a great deal more attention than it's received.

Almost everyone using a quad parasitic array is pleased with the results. The quad has a good capture area, and for modest height above ground, a favorable low angle of radiation. A quad mono-loop has these same advantages, but with the disadvantage of being bidirectional.

My new location precluded a tower and rotary beam. As an alternative I erected thirty feet of tv mast on the chimney of the house; and from the top, some forty feet above ground, suspended a 20-meter quad loop **(fig. 1)***. Because of limitations in guying the sides at the proper height, the antenna looks like an ARRL diamond, but it works like a gem (pun intended)!

Reports broadside to the antenna are generally 6 to 12 dB higher than those received from a three-half-wave doublet (100 feet long, center fed; 40 feet high) favoring the same directions. In fact, except that the antenna is fixed, it performs just about as well as a two-element quad in a former location.

For a 20-meter loop, I suggest starting with sides 17 feet 10 inches long and trimming to optimize the swr. In my case, the regular formula length $250/f_{MHz}$ = side in feet resulted in a resonant frequency higher than desired; probably because the sides are elongated into a diamond shape.

Hams having towers high enough to accommodate a 40-meter loop might find it would outperform a conventional inverted-V doublet, even though the bottom current loop is close to the ground.

James A. Gundry, W8BW

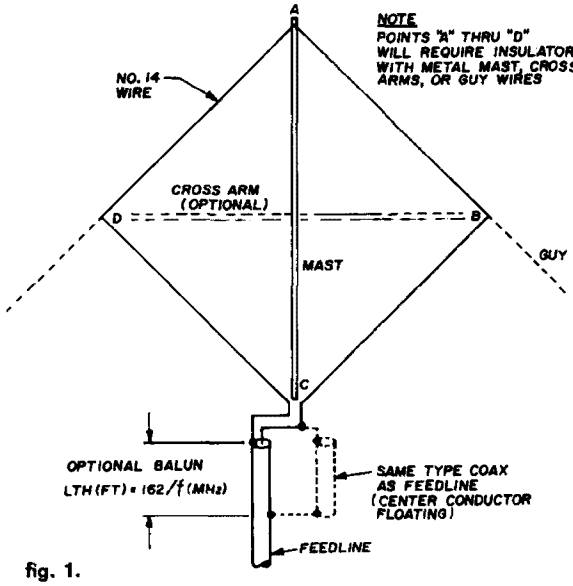


fig. 1.

f (MHz)	total length	length per side
14.0	71′ 10″	17′ 11½″
14.1	71′ 4″	17′ 10″
14.2	70′ 9″	17′ 6¼″
14.3	70′ 3″	17′ 6¾″
21.0	47′ 10″	11′ 11½″
21.1	47′ 8″	11′ 11″
21.2	47′ 5″	11′ 10¼″
21.3	47′ 2″	11′ 9½″
28.0	35′ 10″	8′ 11½″
28.5	35′ 3″	8′ 9¾″
29.0	34′ 8″	8′ 8″
29.5	34′ 1″	8′ 6½″

* **Editor's note:** Suggestions for crossarm braces and a balun are included. The balun is recommended for optimum efficiency.

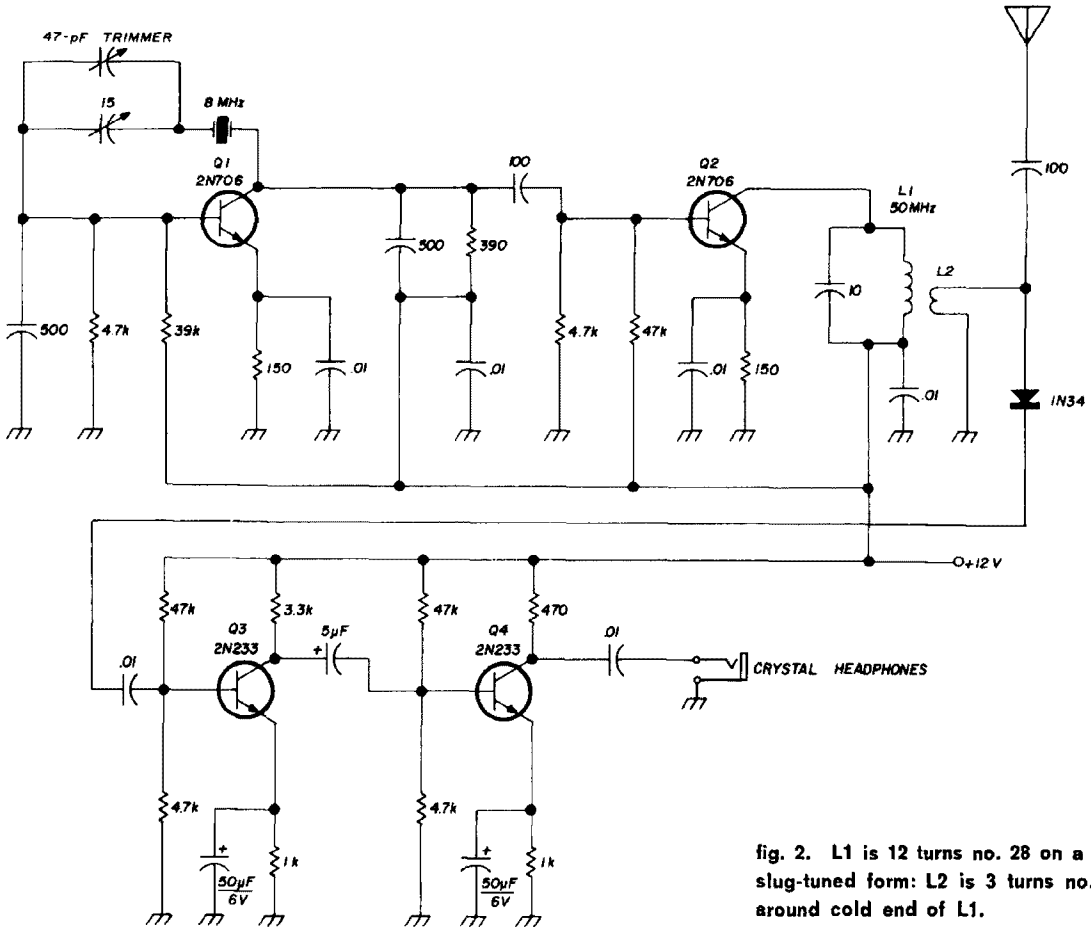


fig. 2. L1 is 12 turns no. 28 on a ¼" slug-tuned form: L2 is 3 turns no. 28 around cold end of L1.

crystal-controlled frequency meter

For the past couple of years amateur radio magazines have been running articles about sophisticated frequency measuring devices, mostly using integrated circuits and other solid-state components. The crystal-controlled frequency meter shown here is as simple as can be constructed and is based on the old method of zero beating one frequency against another.

The oscillator in **fig. 2** is an untuned crystal-controlled oscillator. The crystal frequency can be "pulled" from 2 to 4 kHz, depending on the crystal. I use an 8 MHz crystal and multiply it to 50 MHz; the 50 MHz signal is rectified and fed to an audio amplifier. With this carrier turned on, every time another carrier of the same frequency is fed to the amplifier a zero beat is heard. It even radiates enough to check receiver frequency.

The stability of the oscillator was checked with a digital frequency counter and only ±10 Hz drift was observed at 8 MHz during a two-hour period. The frequency meter uses a 12 volt battery supply; current drain is 8 mA—due chiefly to the low-level of amplifier used in the circuit.

Construction is not critical. It can be hand wired or everything can be put on printed circuit boards; I prefer printed circuit boards because it is quicker as well as compact and neat. Transistors are the four-for-$1 type and crystals were 50¢ each. The transistors in the amplifier should be npn germanium with a beta of at least 90; the 2N233 or 2N170 will do the job fine.

Crystals from 5 to 13 MHz will work in the oscillator. Adjust the 47-pF trimmer to zero a certain frequency with the proper crystal; make sure the 15-pF capacitor is half open. You should be able to deviate the crystal frequency ±1.5 kHz. To set a transmitter vfo, set the 15-pF capacitor to the desired frequency, plug in a crystal headphone, load the transmitter into a dummy load and vary transmitter vfo to zero beat.

John J. Sury, W5JSN

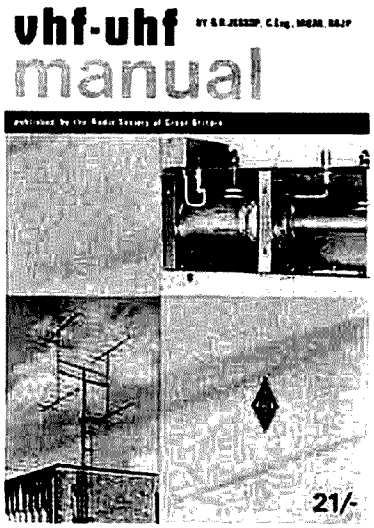
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simple microphone stand

A microphone is usually easy and inexpensive to obtain, but getting a microphone stand is something else again. The few commercially available stands are rather expensive, often costing as much or more than the mike they support. So after looking for

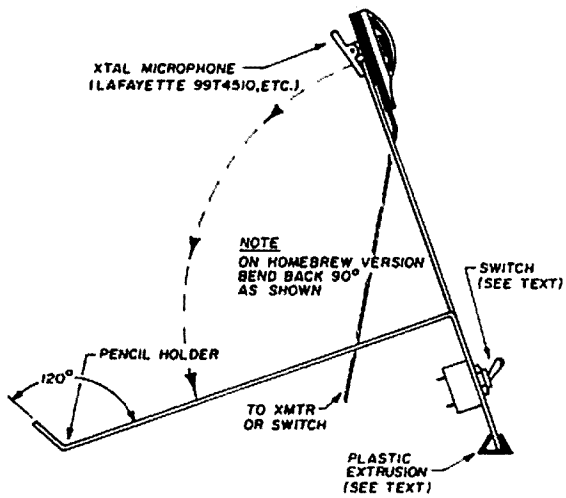


fig. 3. Simple microphone stand.

an easy method of building a stand, I got the idea of using a common, stamped steel bookend turned on its side.

The microphone is attached to the flap of the bookend. The inexpensive, Japanese crystal mikes can be clipped directly to the flap, as they are sold with a built-in clip. Other microphones will require other means of mounting. Epoxy glue should be used if a permanent microphone and stand combination is desired.

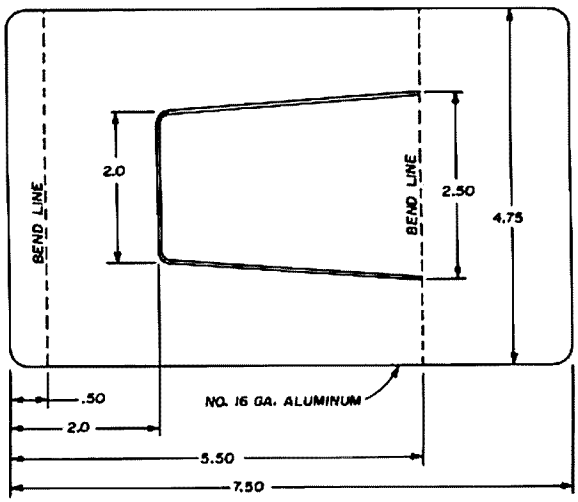
A switch may also be installed on the stand at any convenient location. Details will depend on the type of switching desired (i.e., push-to-talk, etc.).

To prevent marring the surface of your operating desk, attach a rubber or plastic moulding to the bottom of the stand. A good item for this is a piece of the triangular, extruded plastic strips sold for pennies at a stationery store. This material is used for binding loose pages into a report or cata-

* My bookend was purchased in a book store and is made by the W. T. Rogers Company in Madison, Wisconsin.

logue. You may even want to attach a small calendar to the extrusion.

The rear end of the stand can be grooved by bending the steel as shown. This provides a handy place for pen or pencil and prevents scratches as well.



Though a bookend of the type shown* can be bought completely finished and painted for about fifty cents, for those who prefer to make one out of aluminum or steel sheet, a suggested layout is given in fig. 3.

D. E. Hausman, VE3BUE

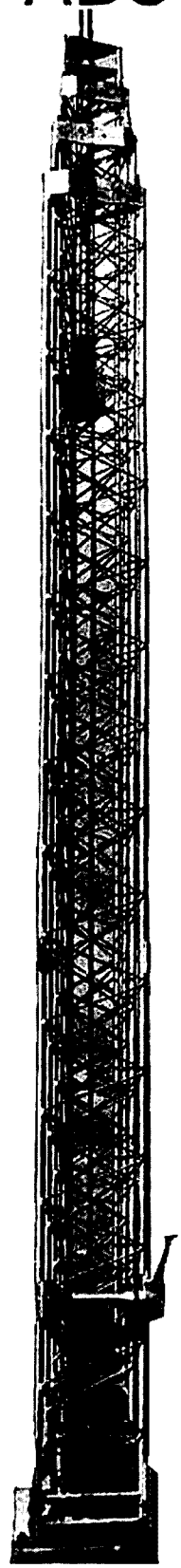
uhf tuner tester for tv sets

Most amateurs have tv sets, so I thought this idea would appeal to **ham radio** readers. It's an adaptation of a miniature solid-state GE tuner for use as a piece of test equipment. With this tester, you can check the uhf front end of your tv set without a signal generator.

A number of surplus-type uhf tuners are available that will work nicely as signal generators. I purchased mine for $5.20 postpaid from Arcturus Electronics.*

* GE transistorized tuner, catalogue number GE 85 Arcturus Electronics, 502 22nd Street, Dept. HR, Union City, New Jersey 07087.

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Remove the tuner from its cover by unhinging the securing wire. Note the junction of two resistors feeding the stator of the oscillator section of the tuning capacitor. Attach a one-half-inch piece of insulated wire to this junction. Drill a hole in the cover to accommodate the wire when the cover is replaced. Leave about one-eighth inch of bare wire at the other end. Replace the cover.

Connect a half-watt, 330-ohm resistor to the insulated stud. (This is the spot where the tuner instructions say to install the 15 kilohm resistor furnished.) Attach a few inches of wire to the other end of the resistor. Tape wire and resistor to the tuner case. Install the mounting bracket and nut according to tuner instructions, as well as the fine and coarse tuning knobs. Connect a 6-volt battery charger (must be unfiltered), with negative lead to the case and positive lead to the wire connected to the 330-ohm resistor.

Tune the tv set to a known uhf channel. Adjust the tuner slightly so the picture just disappears. Run a jumper from the wire emerging from the oscillator section of the tester to the uhf antenna terminals of the tv set. Apply power to the tuner, and tune until two black and two white bars appear on the tv screen. (This assumes a full wave charger or rectifier circuit.) Set the tester dial to read the same as the uhf dial on the tv set minus the offset to make the picture disappear.

This completes the tune-up and tester calibration. Other modulated 6-12 volt sources may be used, providing the transistor in the tester isn't overdriven.

John R. Schuler

transistors for vhf transmitters

Amplitude modulation, although popular with hams on the vhf bands, is not popular with transistors. If you really want to know how to get high quality a-m, get copies of the RCA 40290/40292 transistor data sheets. If a particular rf power transistor is not rated for a-m use, then, "10 watts out," means 2 or 3 watts of carrier power and 10 watts on a-m peaks. For a-m operation with a 12-volt power

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supply use a 28-volt transistor and modulate the last two or three stages.

For single sideband, a 10-watt rf power transistor runs 10 watts PEP—not very attractive. Since transistors heat up in thousandths of a second, they don't "average" very well. Also, there's no such thing as ICAS ratings for rf power transistors.

Type	Power Output (Watts) 6M	2M	220	432	Comment	Price
40405	0.5	0.4	0.4		12-V doubler or amplifier	$ 1.65
2N3866	1.6	1.5	1.3	1.0	28-V driver but good with 12-V supplies	2.97
2N4073	1.0	0.6	0.3		12-V driver	2.70
40290	2.0*	2.0*			rated for 12-V a-m	2.48
40292	6.0*	6.0*	4.0		insulated collector; rated for 12-V a-m	10.45
2N2631	7.5				up to 40-V, poor with 12-V supplies	3.47
2N3553	8	5	3		TO-39 case	4.37
2N3375	10	8	6	2.0	insulated collector	10.80
2N2632	15	13	10		insulated collector	12.75
2N3733	20	20	15	8	insulated collector	15.80
2N5016			25	20		43.50

* amplitude modulated

For fm or CW exciters, rf power transistors perform well. However, if the last stage is driven hard, it will require some sort of protection from mismatched loads —either a directional-coupler scheme or an isolator. High power rf amplifiers that are used to drive varactor frequency multipliers are especially likely to have mismatch problems.

The power gains of the transistors listed in **table 1** are comparable to commonly used vacuum tubes; net efficiencies, including circuit losses, are often greater than 70%. Those listed are ones I consider best buys in terms of results.

Hank Cross, W1OOP

75 cents

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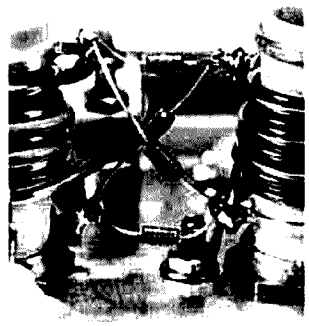
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ham radio

magazine

OCTOBER 1969

HOT CARRIER DIODES



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noise blanker

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staff

editor
James R. Fisk, W1DTY

vhf editor
Nicholas D. Skeer, K1PSR

fm editor
J. Jay O'Brien, W6GDO

associate editors
A. Norman Into, Jr., W1CCZ
Alfred Wilson, W6NIF
James A. Harvey, WA6IAK

art director
Jean Frey

publisher
T. H. Tenney, Jr. W1NLB

offices

Greenville, New Hampshire 03048
Telephone: 603-878-1441

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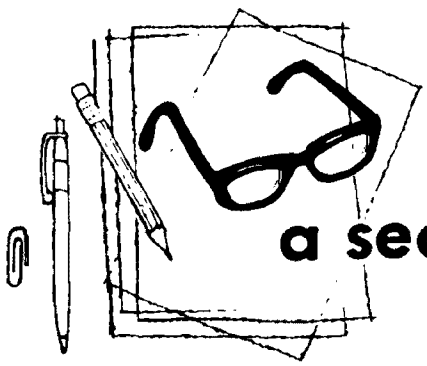
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a second look

by jim fisk

During disastrous hurricane Camille along the Gulf Coast, and the terrible flooding in Virginia, amateur radio was once again called upon to provide communications in and out of the stricken areas. Most of the emergency communications from Mississippi were handled on the high-frequency bands, while two-meter fm was used in Virginia.

The many fm repeaters around the country are ideal for emergency work since they provide dependable long-range communications with low-powered battery-operated transmitters. The Lynchburg, Virginia, repeater, WB4HCX, was tied in with the communications control station K4KJN, and handled communications with the isolated disaster areas. At one point W4GCE's remote-control station was placed into service to handle the heavy load of traffic. After approximately 100 hours of continuous operation, minimum telephone service was restored, and K4KJN, repeater WB4HCX and remote-station W4GCE ceased emergency operation.

Two-meter fm has an additional advantage during an emergency—the communications-control station can exercise better control over the situation. When operations shift to the high-frequency bands, communications are hindered by the many stations who are "standing by" to help. If the stations standing by would do more listening and less transmitting, operations would be speeded immensely. Also, during the recent crisis, there was all too much idle chatter on the emergency channel. Several times emergency and medical traffic was held up while stations broadcast various "announcements" and news reports to the waiting multitude.

The next time you hear an emergency frequency in operation, don't jump in with both feet and announce that you are "standing by" to help unless you're asked to do so. If the net-control station is looking for a station to accept traffic to your vicinity, **that** is your cue to turn on your transmitter. Remember too that amateur stations in the stricken area will seldom handle incoming traffic because they have no way to get it to the addressee.

catalog time

If your mailbox is anything like mine, it's been overloaded of late with new catalogs from all the big electronics distributors. First on the scene was the new 1970 number from Allied Radio, followed closely by volumes from Burstein-Applebee and Lafayette Radio Electronics. The new 1970 Radio Shack catalog is out too, but for that one you have to personally call at your local Radio Shack outlet.

If you do any building at all, you owe it to yourself to have these catalogs in your shack. If you're not already on your favorite distributor's mailing list you probably didn't buy any components during the past year or so, but a post-card request will usually bring a copy of his latest catalog *tout de suite*.

Jim Fisk, W1DTY
editor



hot-carrier diode converter for two meters

Hot-carrier diodes simplify the construction of high performance low-noise converters

Gary Vander Haagen, K8CJU, 145 Enterprise Drive, Ann Arbor, Michigan 48103

In a conventional vhf or uhf converter consisting of an rf stage, oscillator-mixer and output i-f amplifier, the mixer is usually noisy and operates with substantial loss in signal. To overcome these problems, it is necessary to use a low-noise moderate gain front end and an output i-f amplifier to obtain a good noise figure along with over-all converter gain. The converter stage is generally responsible for this increased complexity, and although the problem is slowly being solved with improved semiconductors, one solid-state device that is ideal for this use has been largely ignored by amateurs—the hot-carrier diode.

Until recently the high cost of this device precluded its use in all but prestige military and high priced commercial equipment, but now prices are within the budget of the amateur. The hot-carrier diode is characterized by low-loss frequency conversion and extremely low circuit noise, so it makes an ideal frequency mixer stage.

By using hot-carrier diodes in an amateur converter, the rf amplifier and i-f output stage can be eliminated; all you need

is the hot-carrier diode mixer and local-oscillator injection. A simple two-meter converter using these techniques is described in this article. This converter performs as well as any conventional high-performance unit with only a fraction of the circuitry. Except for the diodes, all the components are the junk-box variety. The total cost of the five diodes is $7.50.

hot-carrier diode mixers

The hot-carrier or Schottky diode is a metal-on-semiconductor diode that rivals the point-contact diode (1N21 for example) in performance but far surpasses it in uniformity, reproducibility and reliability. Current through these devices is primarily by means of majority current carriers; usually electrons since they are the most mobile. When the hot-carrier diode is forward biased, the current carriers are injected into the metallic portion of the junction at energy levels much higher than free electrons, thus the name, "hot carrier".

For frequency-conversion applications the low-noise high-efficiency characteristics of the hot-carrier diodes are especially desirable. Conversion efficiency is typically −2 to −3 dB (ratio of i-f power to incoming rf power). The low noise contributed by the hot-carrier diode is a direct result of the device's characteristic current-vs-voltage curve.

The curves of **fig. 1** compare the characteristics of the point-contact diode with those of a hot-carrier device. It can be seen that the reverse breakdown voltage for the hot-carrier diode is much larger than that for the point-contact diode. With the higher reverse breakdown voltage of the hot carrier diode, it conducts less easily in the reverse direction, so is much quieter at normal operating voltages. The fact that reverse diode current is noisy should be obvious to any vhf experimenter who has used reverse-biased diodes as noise sources.

When a diode is used as a mixer, the local-oscillator signal alternately turns it on and off. When the local oscillator voltage swings in the reverse direction, a large amount of shot noise is generated by the point-contact diode. With the hot-carrier diode, very little noise is generated when the local oscillator voltage swings in the reverse direction.

As noted before, the higher efficiency and lower noise of the hot-carrier diode means that rf and i-f amplifier stages are not required. However, since some loss is introduced by the mixer, there is some reduction in over-all receiving system sensitivity. If, for example, a hot-carrier diode converter is used with a communications receiver that has a sensitivity of 1

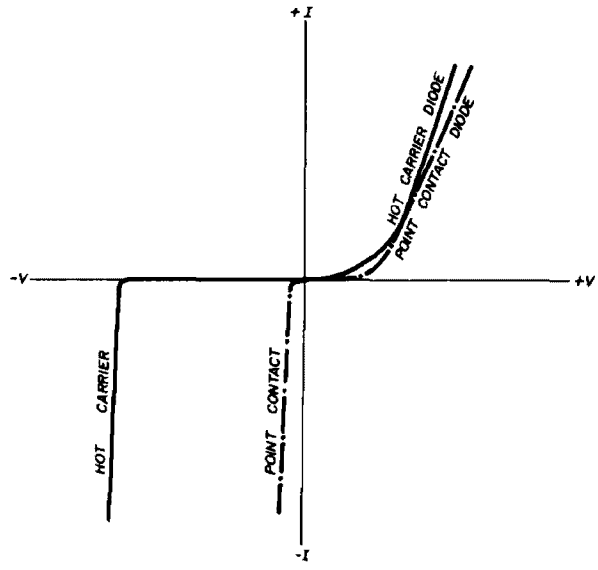


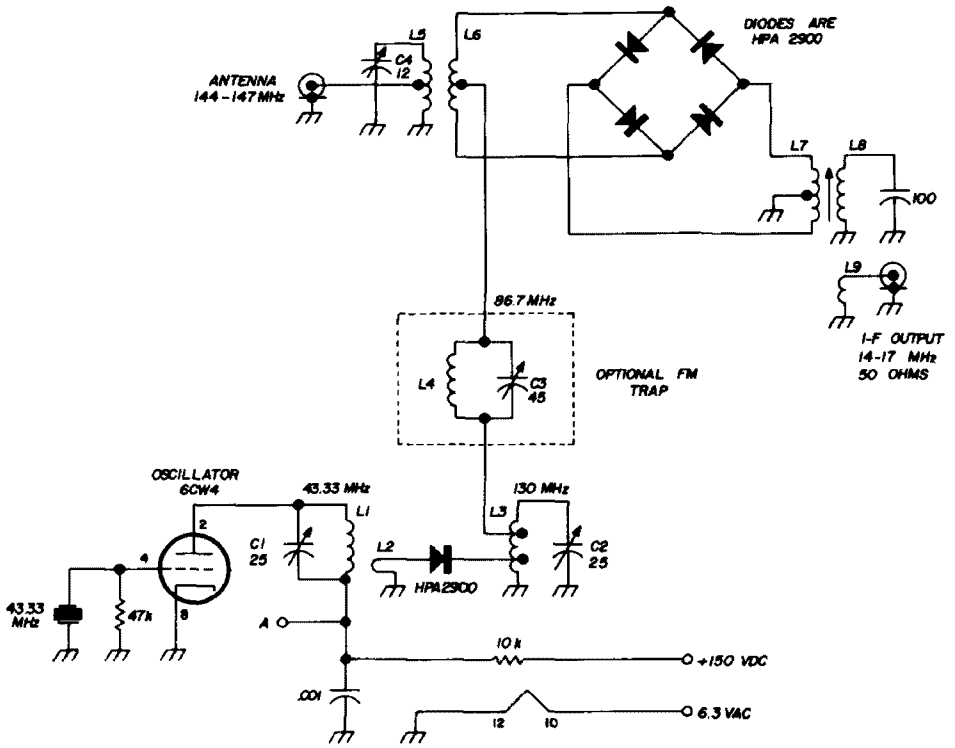
fig. 1. Volt-current characteristics of typical hot-carrier and point-contact diodes.

μV for 10 dB signal-plus-noise-to-noise ratio, the over-all receiving sensitivity with the converter is reduced to 1.6 to 2.0 μV for the same performance. This is still very competitive with existing high-performance converters.

Another characteristic in favor of the hot-carrier device is its large dynamic square-law range which means fewer problems with cross modulation from nearby high-level signals.

two-meter converter

The two-meter converter circuit shown in **fig. 2** consists of a double-balanced hot-carrier diode mixer driven by a crystal-controlled oscillator. The double-balanced



C1, C2	4.5-25 pF padder (CRL 822CN-N650)
C3	7-45 pF padder (CRL 822BN-N650)
C4	3-12 pF padder (CRL 822FZ-NPO)
L1	0.78 μH, 10 turns no. 20 air wound, 1/2" diameter 5/8" long (B&W 3003 coil stock)
L2	1 turn loop wound around bypassed end of L1
L3, L4	0.21 μH, 4 turns no. 18, air wound, 1/2" diameter, 5/8" long (B&W 3002 coilstock) Diode tap 1 1/2 turn from ground end; output tap 2 1/2 turns from ground end.
L5	5 turns no. 20 wound on 3/8" ceramic form, center tapped (J. W. Miller 4400 with slug removed); see fig. 4A.
L6	3 turns no. 20 enamelled wound over L5, center tapped; see fig. 4A.
L7	3 turns no. 20 enamelled wound over L8, center tapped; see fig. 4B.
L8	11 turns no. 20 enamelled close wound on 3/8" slug-tuned form (J. W. Miller 4400-2); see fig. 4B.
L9	2 turn loop wound over ground end of L7, L8.
RFC	2.0 μH, 20 turns no. 26 enamelled close wound on 1/4" form (National R60)

The HPA 2900 diodes may be purchased from any Hewlett Packard Sales office. Consult the Yellow Pages or write to Hewlett Packard, 620 Page Mill Road, Palo Alto, California 94304, for the address of your local sales office.

fig. 2. Schematic of the high-performance hot-carrier diode converter for 144 to 147 MHz. If you want all solid-state construction, use the transistor oscillator circuit shown in fig. 3.

mixer has a tuned input and tuned output and uses four Hewlett-Packard 2900 hot-carrier diodes. A fifth diode is used as a frequency multiplier from 43.33 MHz to 130.0 MHz. Local oscillator power is injected at the center tap of the input inductor, L6. Although a nuvistor local-oscillator stage is shown in **fig. 2**, a transistor stage, such as the one shown in **fig. 3**, could be used as well.

The oscillator circuit uses a third-overtone 43.33 MHz crystal. The output is loop coupled to the hot-carrier diode tripler; the tank circuit L3-C2 is resonant at 130 MHz. An optional fm trap is included in the local-oscillator injection line for use where strong signals from fm broadcast stations may mix with the second harmonic of the oscillator. If no strong fm stations are in the area the fm trap may be eliminated.

construction

The converter is built into a 2 1/8x3x5 1/4 inch minibox. The oscillator coil L1, tripler coil L3 and trap coil L4 are commercial air-wound inductors; the mixer

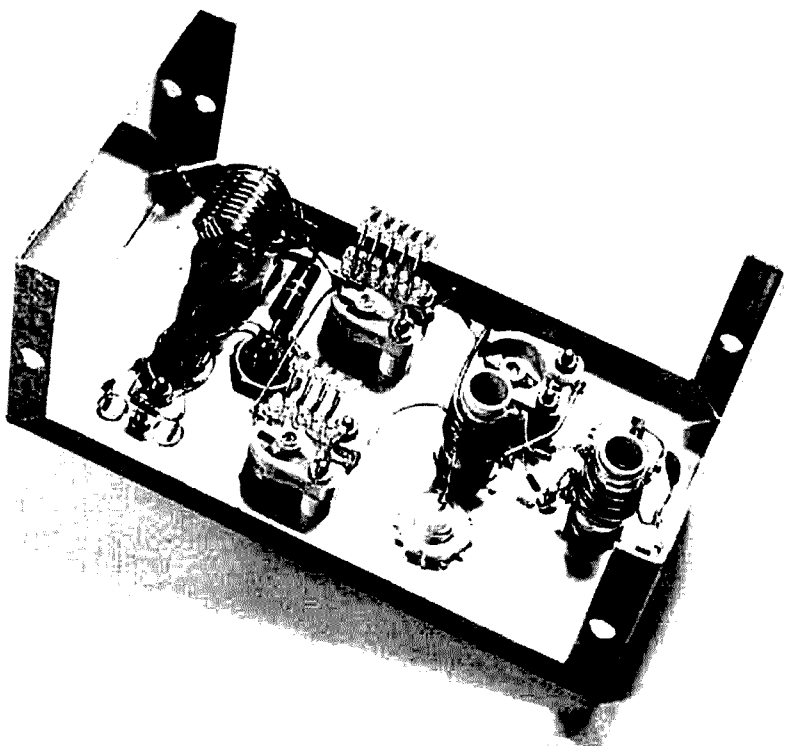
coils are wound on ceramic coil forms. Although the actual layout is not critical, good vhf construction techniques should be followed; all leads should be kept as short as possible, and the coils mounted at right angles for reduced coupling between circuits.

The nuvistor socket is mounted adjacent to the crystal and power socket. The four 45-pF trimmers (C1-C4) are mounted with ceramic spacers placing them about 1/8 inch from the box surface.

For L6 interwind 3 turns of number 20 insulated wire between the turns of L5 and connect the ends to the two coil lugs. Remove a small section of insulation at the center-tap point but be careful not to short the windings; connect an additional number 20 wire at that position. This is the oscillator connection.

The output transformer requires three lugs. If a junk box terminal is not available, you can use a small standoff insulator or terminal strip, but use minimum

Component layout of the two-meter converter. The crystal oscillator and multiplier stages are to the loft; the hot-carrier-diode mixer is on the right.



While the two mixer transformers are not especially critical, care in construction will improve the balance and oscillator rejection in the completed converter. The two transformer assemblies are shown in **fig. 4.** For the input transformer first wind 5 turns number 20 bare copper wire around the form, leaving about 1 1/2 inches at each end. Spread the turns so that the L6 winding may be placed between the turns. The coil should be about 1/2 inch long. Use a short piece of bare number 20 wire for the antenna tap.

lead length. First, wind 11 turns of number 20 insulated wire (L8) on the second form. Terminate one end on the upper coil terminal and leave the other end free for connection later to a ground lug. Secondly, wind 3 turns of number 20 bare wire (L7) over the center portion of L8 and attach both ends to the remaining coil lugs. Use a piece of number 20 wire for the center-tap local-oscillator ground return. The remaining output coupling loop (L9) is wound tightly over the ground end of L8 with the ends left for later connection to

ground and the output connector.

The four hot-carrier diodes should be mounted between the four coil lugs as shown in the photograph. Don't use excessive heat when soldering them into the circuit.

alignment

If the fm trap is used, put a jumper wire across this circuit before beginning the tune-up procedure. With the crystal and nuvistor in place, apply power to the circuit and adjust the oscillator for maximum output. Attach a dc voltmeter between ground and the junction of L1 and the 10k resistor (point A). When the 6CW4 is not oscillating, this voltage should be about 50 Vdc. Adjust capacitor C1 for maximum oscillator output while simultaneously checking that the oscillator will start. This can be accomplished by switching the B+ off and on after each adjustment. With the oscillator properly adjusted, the voltage at A will read 70 to 100 Vdc.

The transistor oscillator in **fig. 3** is adjusted by connecting a dc voltmeter between ground and the rfc–47-ohm resistor-

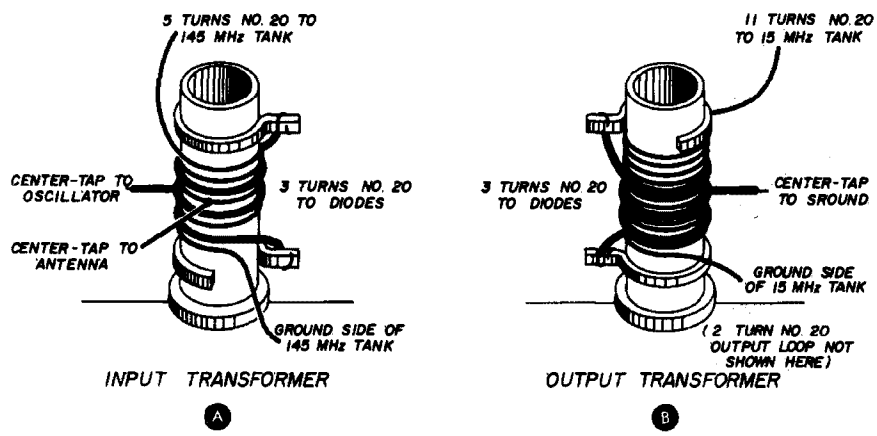
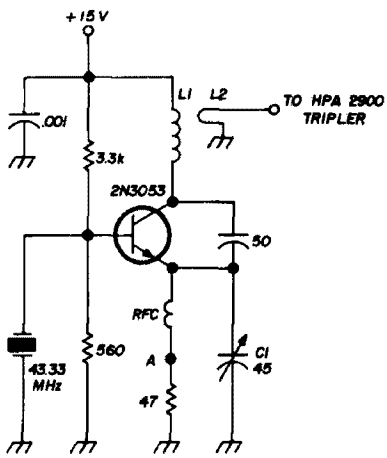


fig. 4. Construction details of the input and output mixer transformers. Each of the transformers is wound on a 3/8" ceramic form such as the J. W. Miller 4400. No tuning slug is used with the input transformer.



C1	**7-45 pF padder (CRL 822BN-N650)**
L1	**0.38 μH, 6 turns no. 20, air wound, ½" diameter, 7/16" long (B&W 3003 coilstock)**
L2	**1 turn loop over bypassed end of L1**
RFC	**2.0 μH, 20 turns no. 26 enamelled close wound on ¼" form (National R60)**

fig. 3. Third-overtone transistor oscillator circuit suitable for the two-meter converter shown in fig. 2.

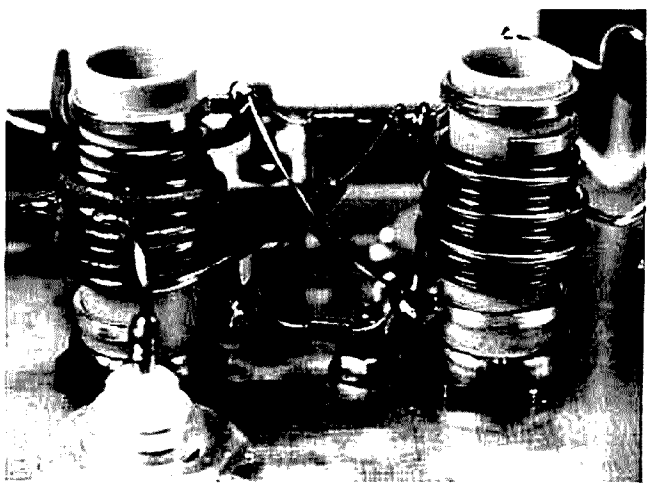
junction (point A). When the device is not oscillating, the voltage should read approximately 1.0 Vdc with a 15-V supply. Adjust capacitor C1 for maximum voltage reading (3 to 4 Vdc) and check for oscillator starting.

Connect the converter to a suitable receiver with a short piece of coaxial cable. With a 145-MHz signal applied to the input and the receiver tuned to the correct i-f frequency (15 MHz), adjust the tripler capacitor C2 for maximum signal strength. Then adjust the input circuit C4 for maximum signal strength. Adjust the position of the slug in the output transformer for maximum signal strength. Repeat the oscillator, tripler, input and output tuning steps, this time for maximum signal-to-noise ratio.

If fm interference is present, remove the short across L4-C3 and tune the trap for

minimum interference. Check to see that this adjustment didn't reduce the signal-to-noise ratio. If necessary, repeat the previous alignment steps to negate any interaction that may have resulted.

Local oscillator power should be more than sufficient, but the oscillator loop coupling should be changed (moved or more turns added) to verify that there is no increase in the signal-to-noise ratio as a result of increased oscillator drive. The converter requires 1 to 5 milliwatts of rf drive for optimum performance.



Closoup of tha two mixer transformers and the four hot-carrier diodes.

closing thoughts

Some difficulty generally arises with any oscillator-multiplier system because of subsequent harmonic mixing: mixing of signals with the oscillator harmonics and falling within the i-f band. The optional trap circuit in **fig. 2** should handle most situations where very powerful stations are not next door. However, if fm stations continue through with the trap in place and the circuit is well shielded, about the only surefire alternative is to change the i-f. A double-tuned trap may also help.

The double-balanced mixer circuit and HPA 2900 hot-carrier diodes are suitable for operation over all amateur bands from 160 meters through 1000 Mc with suitable circuit modification.

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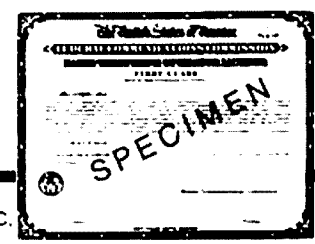
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a practical discussion of product detector operation

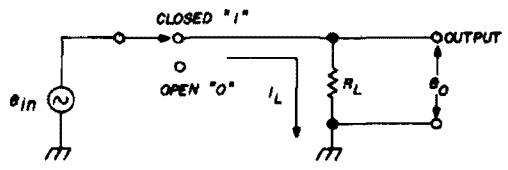
Including a brand new high performance product detector using hot-carrier diodes

Mike Goldstein, VE3GFN, 22 Kingswood Road, Toronto 13, Canada

In the world of ham circuitry, where almost everything these days can be reduced down to oscillators and linear amplifiers, the product detector is probably the last frontier of mystery. The reason is elementary—even the simplest product-detector circuits operate on some pretty high-flying principles. Without delving into some frightening algebra these principles are a little hard for nonmathematicians to grasp. Let's discuss the whole thing down at the grassroots level, so the circuit that follows can be viewed with the utmost appreciation.

To start, the term "product detector" is very misleading. We know that the output of such a detector is basically the sum and difference of the two input frequencies (i-f and bfo), so why is the beast called a product detector? If you can't answer that question without peeking, then it isn't cricket to flip the page and look at the ads.

fig. 1. Output of this simple mechanical switching circuit is the product of the load resistance and the input current.



operation

Let's consider the operation of a simple switch, that is opening and closing, allowing current to flow in some load resistance

(**fig. 1**). When the switch is closed the current caused by the input signal produces a voltage drop across the load resistor which appears as a sine-wave output voltage. This output voltage has an amplitude $E_o = I_L R_L$ by Ohm's law. When the switch is open no current flows in the load resistor, and the output point sits at ground level, or zero volts.

The waveforms of this circuit could be plotted like those of **fig. 3.** If we close the switch at time T1, open it at T2, etc., we appear to generate a square wave having the levels "switches open" and "switches closed." Call the level when the switch is closed 1, and the level when the switch is open 0. When the switch is at 1 the current is I_L; when the switch is at 0 the current is zero. Another way of writing this is:

$$I_L \times 1 = I_L$$

$$I_L \times 0 = 0$$

Note that switching between the two levels effectively generates a square-wave of current. Now, let's bravely write the equations that describe the output voltage for the waveform of **fig. 3,** for the conditions of the switch being open and closed.

T1 → T2:

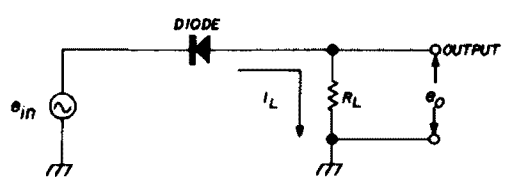
$E_o = I_L R_L$, $I_L \times 1 = I_L$;
so E_o does equal $I_L R_L$

T2 → T3:

$E_o = I_L R_L$, $I_L \times 0 = 0$; so E_o equals zero.

Note that each time we write the equation for the output voltage (E_o), it depends on the product of the value of R_L (which is constant) and whatever amplitude I_L has at the time. Another way of saying this is that the output voltage $I_L R_L$ is multiplied by the square wave that controls the current flow.

fig. 2. Putting a diode in the simple circuit of fig. 1.



doing it electronically

Now let's get rid of the mechanical switch; use a good, fast, silicon switching diode instead. If we examine the characteristics of a switch and a fast diode, it appears that this substitution is legal. Could ten thousand computer engineers be wrong? **Fig. 4** shows the two characteristics.

Note that the silicon diode has a very little voltage drop across it when it is forward biased for conduction. This, of course, is the same with a switch. The switch does not delay in going from minimum conduction to maximum conduction, and the diode doesn't lag far behind. With the diode replacing the switch, let's look at our circuit again (**fig. 2**).

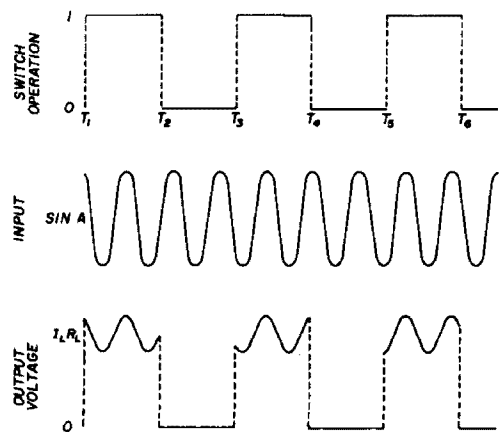


fig. 3. Timing diagram for the circuit in fig. 1 shows output is obtained only when the switch is closed, or in the "1" position.

We cannot close the diode switch by opening and closing a lever, but we can provide an electronic lever in the form of a square wave to do the switching for us. Let's define the square wave as having two levels, 0 and 1. **Fig. 5** shows the diode switch circuit being operated by this square wave.

What happens in the circuit? At T1, the square wave changes from the 0 to the 1 level, charging C to this level. The diode is now forward biased, so the switch has closed. The square wave allows the input voltage to produce a voltage drop across R_L such that $E_o = I_L R_L$.

You can now plot the waveforms for our new circuit. Looking at these waveforms, and comparing them with the waveforms plotted for the simple mechanical switch circuit, it is obvious that the waveforms, and therefore the equations describing those waveforms, are identical for either circuit. We are left, as before, with an output that is the product of the output voltage $I_L R_L$ and the square wave that controls the current flow. The only difference is that now the square wave is electronic, not mechanical. In each case, however,

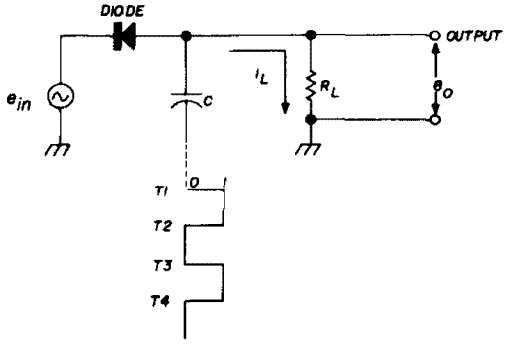


fig. 5. Practical diode switching circuit does the same job as the mechanical switching circuit in fig. 1; output waveforms of this circuit are shown in fig. 2.

the output is the product of the equations of the two input waveforms, so both circuits are detectors.

If we describe the input sine wave voltage as (sin A), and the switching square wave as (cos B + cos 3B + cos 5B + . . .)*, and we multiply these waveforms by trigonometrical methods, we will be presented with the terms (A + B) and (A — B),

*Square waves are made up from an infinite number of odd harmonics, hence the use of the terms 1, 3, 5.

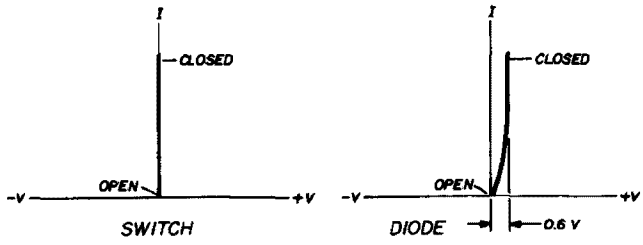


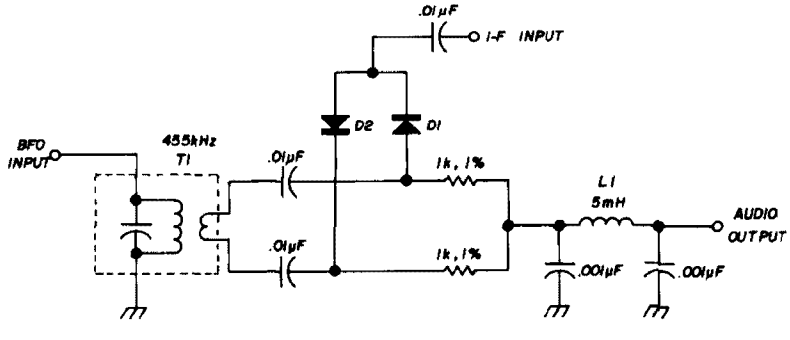
fig. 4. Turn on characteristics of silicon diodes compare favorably with mechanical switches.

among others. These are the sum and difference frequencies we are looking for, and the desired one is easily selected with a properly designed filter. For our purposes, we want only the difference frequency, so we can put a lowpass filter on the output of the detector to reject the high-frequency terms. Note that in perfect multiplication the fundamental input frequencies don't appear in the output.

diode switches

At this point you should understand how and why product detectors work and where they get their name. Look again at the basic diode switch, with a view to developing a working circuit from it. First, consider the output. We said that for a perfect multiplication, no input or switching waveform frequency terms appear in the output. Unfortunately, our diode isn't a perfect switch so this isn't necessarily true. Also, our switch is only conducting over half of the total period of the switching waveform. For optimum operation we have to find some diodes that look most like perfect switches, and make our circuit conduct for the total time, while retaining the multiplying feature of the operation.

fig. 6. High-performance product detector uses two Hewlett-Packard 2800 hot-carrier diodes at D1 and D2 (H-P also manufactures a matched pair, the 2804). The bfo input level should be an 8-volt p-p square wave; i-f input level is 0.3 volts maximum. L1 and the two .001 capacitors comprise a low-pass audio filter.



The way to beat the half-time conduction problem is to have two diodes in the switch such that one conducts for half the switching cycle while the other conducts for the remainder. The problem of the best diode is not solved easily, but a pair of matched hot-carrier diodes is a step in the right direction. These diodes are very fast, can be matched very closely in their characteristics, and have little voltage drop across them when conducting.

One more point that should be examined while we are still in the ivory tower section of this discussion is the subject of the switching waveform. It was stated that a square switching waveform was necessary, and for the best distortion-free output, this is true. However, it is quite in order to use a large-amplitude sine wave to perform the switching function with a good output filter to reject the resulting high-frequency products. Your ear is not likely to notice the difference.

practical circuit

We can now creep out from our ivy-covered hideout and examine a working (boy, does it work) circuit (**fig. 6**) with full understanding. The diodes used are matched hot-carrier diodes produced by Hewlett-Packard, who make wild claims for them, justifiably. The output resistors are precision toleranced to preserve circuit balance. The bfo voltage is fed to the diodes in push-pull to ensure that one is off while the other is conducting.

While the best output waveform was obtained using a square switching waveform, most of us would not bother making a squarewave bfo, and a sine wave switching voltage will produce a good linear output which is clean of hash after putting it through the filter. The audio output level is approximately half the amplitude of the input i-f level; this is only 6 dB attenuation. This circuit compares favorably with circuits using beam-deflection tubes and seems to operate completely within the limits of the theories discussed.

Confidentially, it's the best detector I've seen since I sold my tube puller!

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hot-carrier diode noise blanker

A circuit that provides controlled limiting without distortion— featuring the HP-2800 hot-carrier diode

Ralph W. Campbell, W4KAE, 316 Mariemont Drive, Lexington, Kentucky 40505

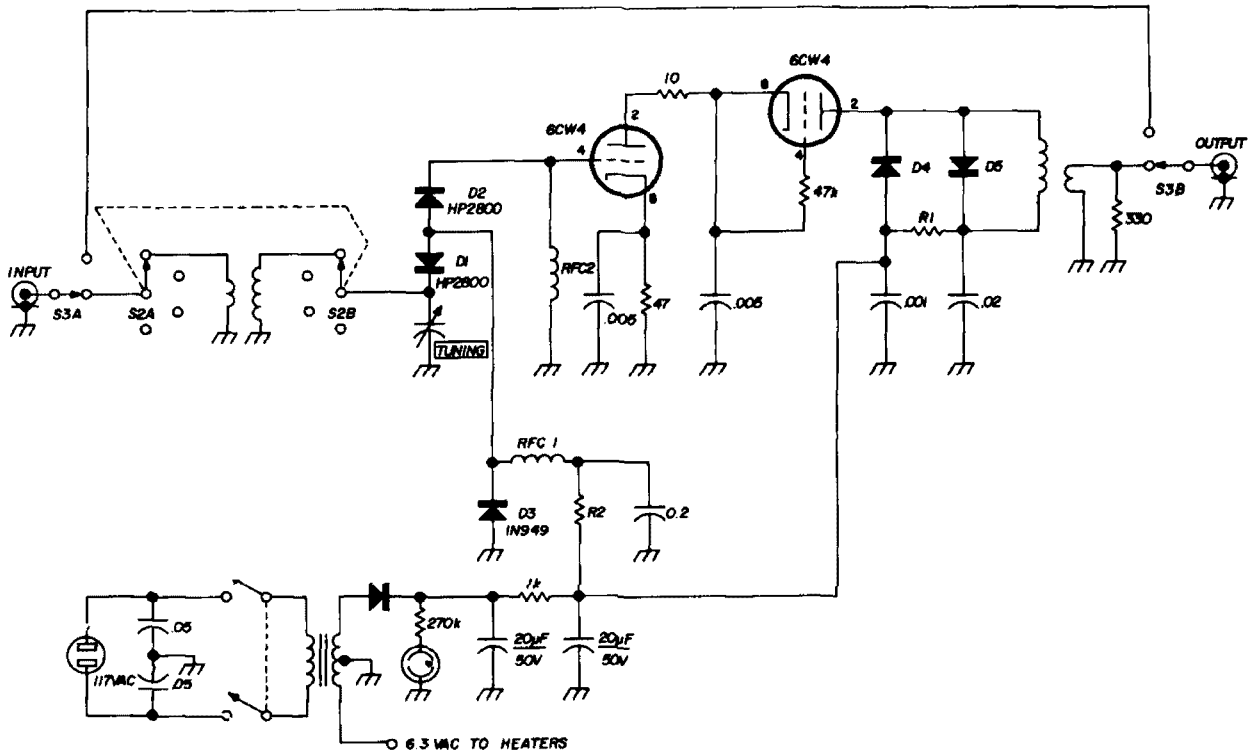
When Hewlett-Packard announced the availability of inexpensive HP-2800 hot-carrier diodes (hcd), I was prompted to find uses for them as replacements for the silicon computer diodes so common in amateur equipment. Chief among their desirable characteristics is near-perfect linearity, not common in silicon or germanium units. The hcd's are closest to diodes such as the Sylvania 1N3731, but they have improved voltage-current characteristics when properly biased. They're also usable in vswr instruments because of their linearity. As mixers, hcd's exhibit comparatively low noise, and they're almost independent of local oscillator injection level. This is important in homebrewed converters without sophisticated equipment.

hcd noise blanker

Using HP-2800's in an i-f noise blanker seemed an ideal application, because they're essentially anti-overload devices. This may seem contradictory, because the usual blanker limits by overloading; however, cross-modulation can be so bad you may as well omit the blanker on local signals. Where these new diodes help is in a trough limiter circuit.[1,2] I adapted this principle to an Ameco PCL-P preamplifier (**fig. 1**) and added a full-wave, shunt-fed voltage doubler that short circuits on high negative spikes. This results in quite precise limiting, which is also effective on auto ignition noise. Output stage overload is prevented by the series input diodes biased to conduct the i-f signal and negative-going spikes only. Signal-to-noise ratio is greatly improved; in fact, with

the avc off you can see the blanker "ride back" on the random noise.

I found it necessary to build up incoming signal plus noise to about 20 dB so there would be adequate voltage threshold for noise suppression. This is the reason for building the blanking circuit into the PCL-P. I substituted 6DS4's for the sharp cutoff 6CW4 nuvistors to control overload on local signals. A big advantage in using the PCL-P adapted to a blanking circuit is that you can tune to any i-f range desired.



D1, D2, D3	Hewlett-Packard HP-2800	R1	68 ohms, 1/2 watt, 5%
D5	1N949, 1N995, or 1N3125	R2	150k, 1/2 watt, 10%
D4	HP-2800 or 1N3125	RFC1, RFC2	75 μH

fig. 1. Ameco preamplifier with W4KAE's noise blanker installed. D1, D2 are biased to conduct i-f signals plus negative-going noise pulses, which are clamped by trough diode D3; amplified spikes are then shorted by D4, providing controlled limiting. D3 may be elminated if D4 functions adequately without it.

theory

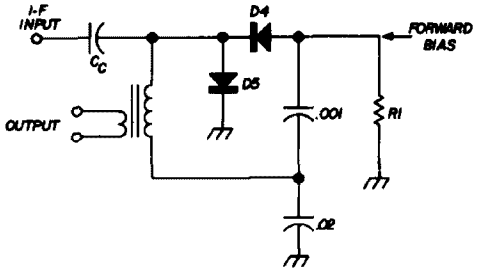
I tried several circuits, but none was as effective as the blanker presented here. The series noise gate was closest to providing desired characteristics, so I improved it to eliminate impulse-noise blasts by adding the trough diode (D3, **fig. 1**). I couldn't eliminate **all** the impulse noise, so I changed the 1N456 silicon junction diode in my SX-122 automatic noise limiter to an HP-2800. Now I can leave the anl switch in the "on" position with no audio distortion.

Assuming high-amplitude signals, here's how the hcd's and the germanium diode control the noise peaks: negative peaks are passed by D1 and D2; amplitude is about +0.3 volt average. The maximum negative-going i-f signal is sharply cut off to 0.22 volt by the trough diode. This means that the total pulse swing in the tee section is from +0.7 to −0.2 volt, with the reference voltage at about +0.3 volt. Leaving the tee network, this reference is clamped to zero volts dc, and only negative-going pulses are passed. The zero-voltage-clamped pulses and noise fragments are passed on the linear characteristic of the forward-biased hcd's; hence there's no loading in the following stage.

Diodes D4 and D5 comprise what I call a "shorted full-wave voltage doubler," because that's how I derived the network. It works! I've redrawn the circuit to explain its operation **(fig. 2).**

You'll see (**fig. 1**) that the output transformer's primary is heavily loaded; its secondary impedance is so low it's impractical to shunt limit here. In **fig. 2** I've shown D5's cathode and the lower end of R1 at ground potential, for illustration purposes. Thus R1

fig. 2. Derivation of the shunt limiter. D4, D5 comprise a "shorted" full-wave voltage doubler; in actual circuit R1 provides bias for D4.



acts as a load resistor. D4 is in series with this hypothetical load and is shunted by the 0.001 and 0.02 μF capacitors. The i-f transformer return is connected to their junction.

Thus, if we have no need to extract power from such an arrangement, D5 can be replaced directly across the transformer primary. When conducting on noise peaks, D4 short circuits negative-going noise pulse since it's forward biased. R1 now serves as a bias resistor **(fig. 1).**

the complete circuit

Typical resistor values for positive diode bias input gating and output loading are shown in **fig. 1.** I replaced Ameco part designations with those shown in parentheses. You could substitute silicon epitaxial planar computer diodes for the hcd's, but you may have to change the bias resistors. R1, the output diode bias resistor, should be 33 ohms, 5 percent. Forward bias is 0.225 V. D4 and D5 are connected as a "shorted full-wave doubler," as discussed above. R1 should be at least 18 ohms; preferably 33 ohms, 5 percent tolerance. Any resistance higher than 33 ohms will drastically reduce the output.

D3 is the trough diode used with the HP-2800 hcd's. Type 1N949, 1N995 or 1N3125's are preferred. D1 and D2 are the passing/clamping diodes that limit signal plus trough noise through the nuvistors. The amplified, negative-going spikes are then limited in a controlled manner by D4. Bias developed across R2 is about 0.33 volt. A 10-percent tolerance resistor is satisfactory for R2, since its value isn't as critical as R1's.

construction

Because the blanker is built from the Ameco preamplifier, little need be said regarding construction. I made liberal use of button micas capacitors, which are mainly for tie points. Bypassing with large values of disc ceramics is necessary.

If you wish to try a saturation limiter, you can add a 10 kilohm wirewound potentiometer in the **"aux pot"** hole of the preamplifier.

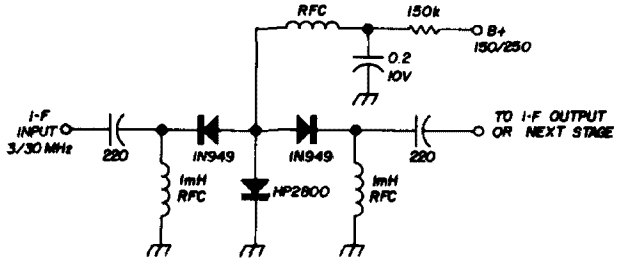


fig. 3. Typical values for the W4KAE Tee noise gate that can be used in usual i-f circuits.

generalized schematic

In **fig. 3** I've included a schematic showing typical values for the hcd blanker that can be added to most receiver i-f strips. If your i-f strip has about 20 dB gain, as mentioned previously, you can add the shunt limiter, using the component values shown in **fig. 1.**

references

1. James Kyle, K5JKX, "Receivers," Radio Bookshop, pp. 37-38.

2. Paul Franson, WA1CCH, "Diode Handbook," *73*, January, 1968, pp. 10A-11A.

ham radio

low-cost integrated circuits for amateur equipment

Simple new consumer ICs should appeal to the home builder who is looking for higher performance with less complexity

Paul Franson, WA7KRE, 7430 E. Minnezona Avenue, Scottsdale, Arizona 85251

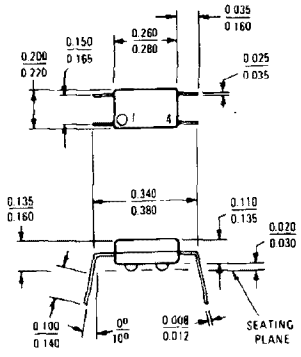
Motorola's recent introduction of simple integrated circuits designed specifically for inexpensive consumer equipment should interest the building ham. Consumer electronics components have always been inexpensive, both because of the huge quantity sold, and because of the demands of equipment manufacturers for lowest costs. But these new consumer circuits are even intended to compete for sockets in the toughest markets of all—those for pocket radios and cassette recorders in the Orient.

It's obvious that this places considerable demands on the American semiconductor manufacturer. The products must offer prices and performance comparable to those obtained from low-cost transistors. This creates quite a challenge to IC manufacturers who have been making parts mostly for industrial, computer and military equipment. However, U.S. manufacturers have considerable experience in the Far East. Most major producers have plants in Korea, Taiwan, Japan or Hong Kong. Motorola's plant in Seoul, Korea, for example, assembles transistors and IC's using completed silicon dice supplied from Phoenix, then ships the transistors back to Arizona for testing and quality control. And plastic transistors manufactured wholly in Phoenix compete aggressively in such penny-conscious areas as Hong Kong. The highly efficient sophistication of mechanized assembly can compete with wage costs far lower than in the U.S.A.

The producers of plastic transistors at Motorola have applied the knowledge gained in producing hundreds of millions of devices to IC production. They are using the same efficient mass-production techniques and automated computer testing used for transistors for making simple, easy-to-use IC's. Instead of 12 or 14 pins, these new IC's have only 4 to 8. The pins are widely spaced (0.16 inch) compared to conventional dual inline packages (DIPs) (0.10 inch) for non-critical circuit-board construction and assembly, yet the packages (**fig. 1**) are a fraction of the size of DIPs to save space (0.20 inch wide in-

stead of 0.72 inch).

A big advantage of these IC's is the simple external connections. Instead of making the circuits from the exceptionally high-performance transistors necessary for very wideband service, the transistors are designed for specific uses—audio or moderately low frequencies. Thus they don't need extensive compensation to prevent oscillation or for desired performance.



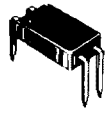


fig. 1. Case outline and drawing of the package used for the new MFC4000 and MFC4010 consumer integrated circuits.

Now let's take a look at two of the new circuits and some of their applications.

audio amplifier

The Motorola MFC4000 is a ¼-watt audio amplifier requiring no transformers. That doesn't sound like much, but it's plenty for most ham applications. Remember that those high-power stereo amplifiers are generally designed for use with speakers having low efficiency, and must be able to deliver extensive power at the low frequencies where our ears aren't very sensitive. The average portable radio has an output of ¼ watt or less, and is more than adequately loud over its restricted bandwidth: the communications bandwidth is (supposedly) even less.

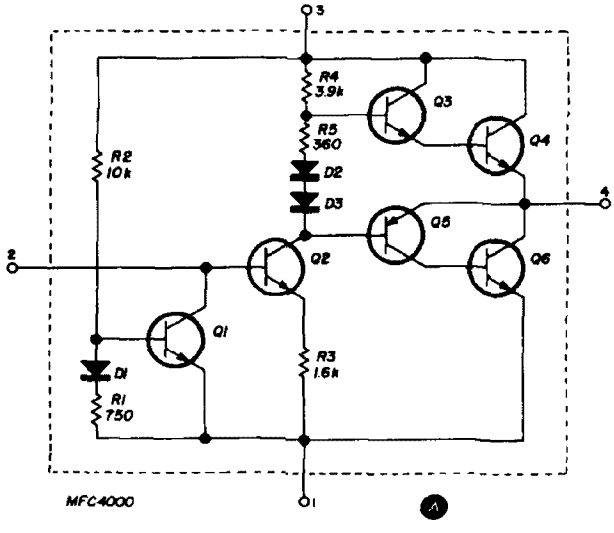
The MFC4000, which costs only $2.10 in small quantities, operates in class B, so idle current is low, typically only 4 mA. Harmonic distortion is low, too, typically under 1%.

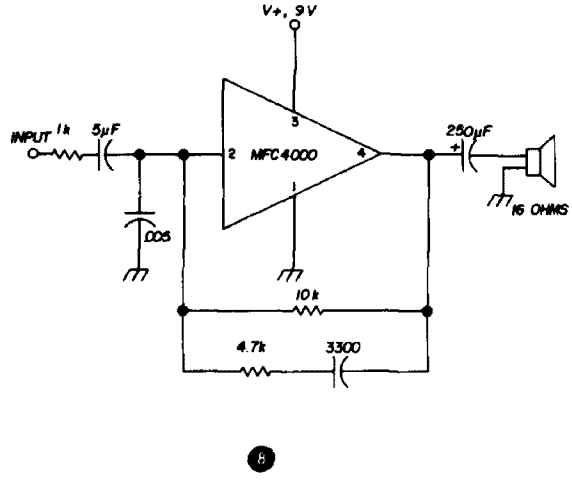
The circuit of the MFC4000 is shown in **fig. 2.** The external test circuit the manufacturer uses is also shown. The IC uses 6 transistors, 3 diodes and 5 resistors. The output is a quasi-complementary circuit consisting of two transistors (Q4 and Q6) driven by a complementary pair (Q3 and Q4). Resistors R4 and R5, with diodes D2 and D3, set and stabilize the bias for the output stages. Transistor Q2 is a driver, while Q1 and its associated components maintain stable bias for the whole circuit.

The external 250-μF capacitor furnishes blocking to prevent shorting the output and to keep dc out of the speaker. The 10-kilohm resistor sets the dc bias and provides dc and ac negative feedback, reducing distortion and providing good thermal stability. The resistor-capacitor network increases feedback at higher frequencies, shaping frequency response.

The .005-μF capacitor at the input stabi-

fig. 2. Circuit of MFC4000 ¼-watt audio amplifier (A) and manufacturer's test circuit (B).





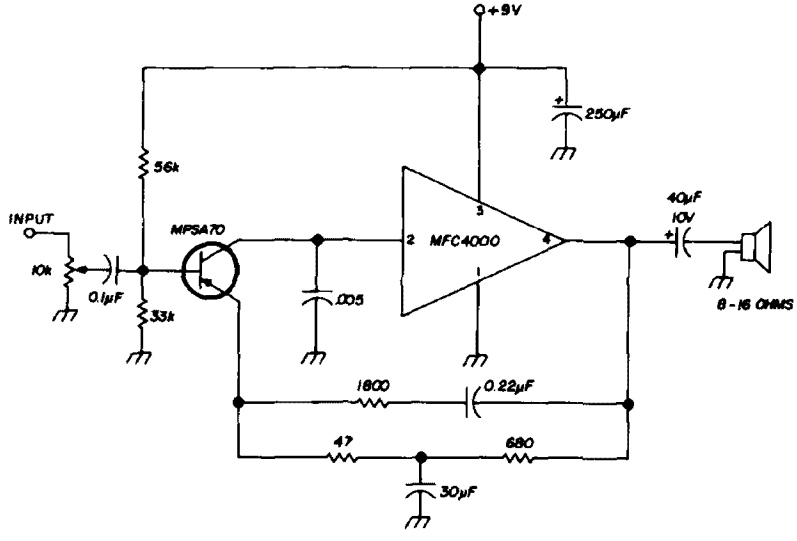
lizes the circuit at high frequencies and prevents reception of local a-m broadcast stations and stray electrostatic fields. The series input resistor is not needed with moderately high impedance signal sources.

This circuit works well. With a 9-V supply, it requires a maximum of 15 mV input for 50-mW out. Typical output power at 10% total harmonic distortion is 350 mW. At more usual output levels (50 mW), distortion is only 0.7%. Maximum power supply voltage is 12 V.

A few words about mounting the MFC4000: no sockets seem to be available, but none are really necessary. Its pins fit perfectly in Vectorboard pattern F. Small rivets provide simple soldering, or the old wrap-around technique is satisfactory. However, some means of dissipating heat is necessary for full output. The maximum power dissipation is 1 watt at 25° (ambient temperature) and should be derated 10 mW/°C above 25° C. For breadboarding, I attached an alligator clip directly to pin 1 (ground) since most of the heat is transmitted through this pin, and the package stayed cool. A better approach would be to use a copperclad circuit board with as much copper area as possible within a couple of inches of pin 1.

Another circuit using the MFC4000 is shown in **fig. 3.** Here an inexpensive transistor, the MPSA70, has been added for high amplification. The feedback resistor has been modified for best results, and the compensation and series capacitors chosen for the limited bandwidth desirable in communications equipment. If you want to widen the bandpass for use in music, increase the input and output capacitors, and the 30-μF parallel feedback capacitor, and reduce the size of the 0.22-μF capacitor.

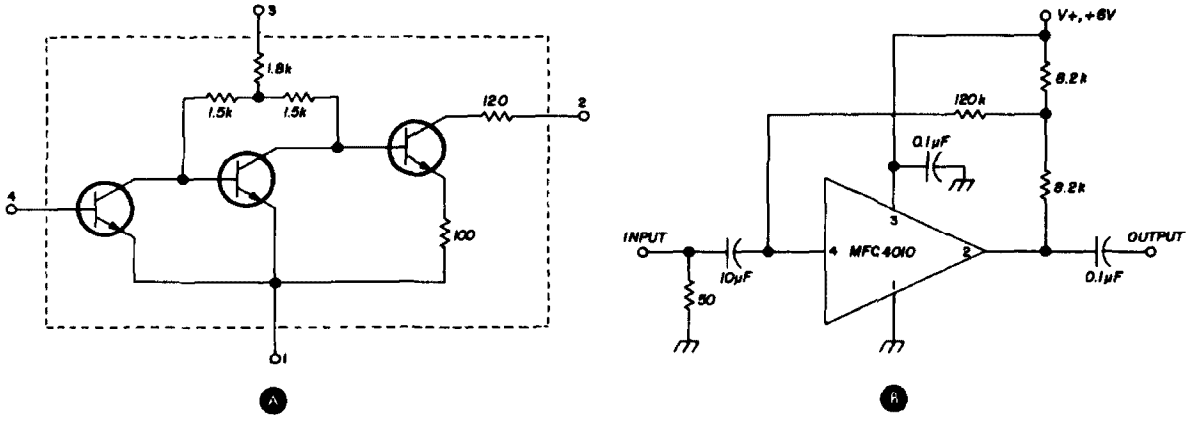
fig. 3. High-gain amplifier with bandwidth restricted for communications applications.



wideband amplifier

The Motorola MFC4010 is a wideband amplifier, or gain block, designed for general audio or i-f applications up to 500 kHz or so. It is a simple, three-transistor circuit providing a minimum of 60 dB gain, and is useful for saving space and components in many applications. The MFC4010, which costs about $1.85, uses the

fig. 4. Circuit of the MFC4010 gain block (A) and test circuit (B). The 120k feedback resistor is relatively critical; the value shown is for 6 V operation.



same small 4-pin package as the MFC4000 (**fig. 1**).

The internal schematic of the MFC4010 is shown in **fig. 4** with its test circuit. As you can see, few external components are required to provide high amplification,

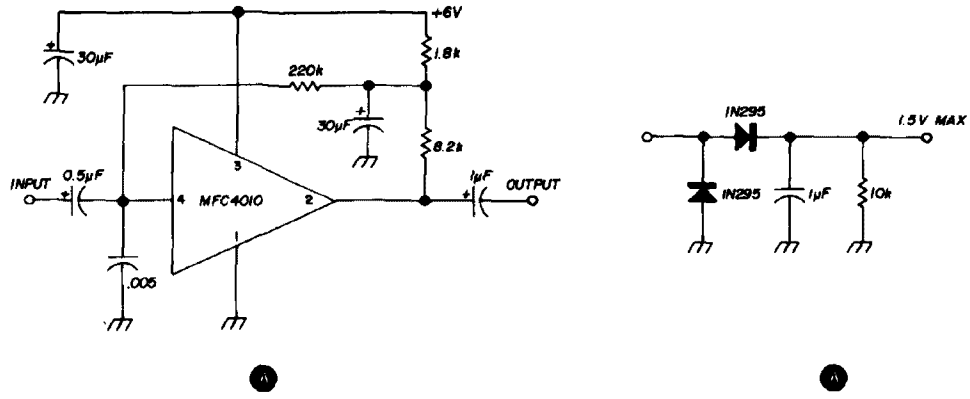


fig. 5. High-gain preamplifier and rectifier for ssb agc, vox or audio compression.

typically 70 dB. Other parameters of interest are maximum power supply voltage of 18 V, maximum power dissipation of 0.5 W at 25° C ambient temperature, and typical current drain of 3 mA. Parameters at 1 kHz are h_{11} of 1000 ohms (equivalent to h_{ie} of a simple transistor), h_{12} (h_{re}) of 10^{-6}, h_{21} (h_{fe}) of 1000 and h_{22} (h_{oe}) of 10^{-5} mho.

Other circuits using the MFC4010 are shown in **fig. 5** and **6.** The high-gain audio amplifier in **fig. 5** is suitable for use as a microphone preamplifier. As in **fig. 3,** capacitor values have been chosen for communications applications. The .005-μF input capacitor is necessary to prevent detection of close-by a-m broadcast stations. A simple voltage-doubling rectifier is also shown. This can be added to the amplifier for use in ssb agc, or vox, or the output can be fed to a controlled dual-gate mosfet for audio compression.

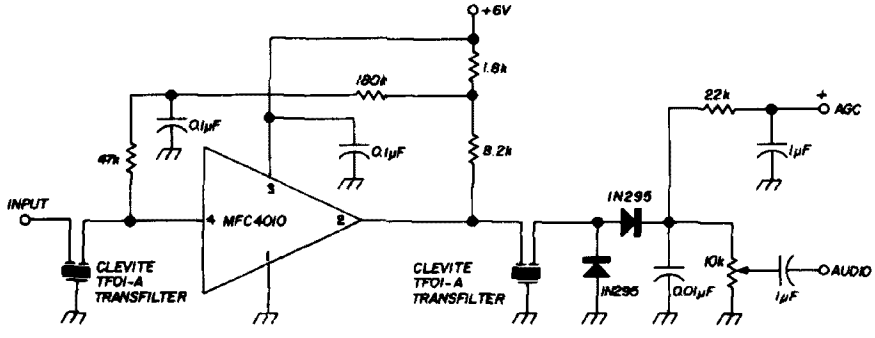
Another interesting application of the MFC4010 is shown in **fig. 6.** This is a complete 455-kHz i-f strip using ceramic transfilters and voltage-doubling detection. It works well with a conventional front end. Agc voltage is provided for controlling stages before the MFC4010. A dual-gate mosfet (MFE3006) would be ideal as a mixer or rf amplifier. The transfilters could be replaced by a mechanical or crystal filter, or by more conventional i-f transformers. Don't forget to provide dc blocking for the input and output if you use a device without blocking (the transfilter appears as a capacitor).

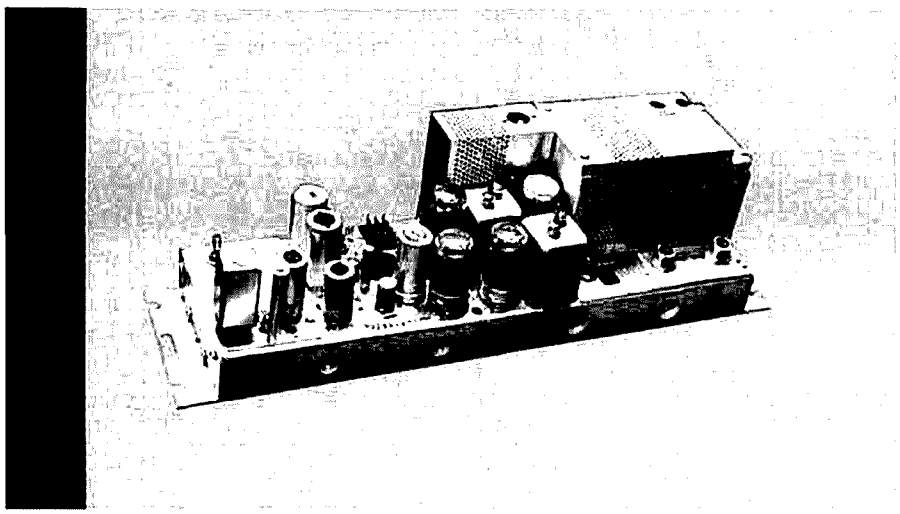
summary

New consumer integrated circuits such as the Motorola MFC4000, and MFC4010 provide many possibilities for the ham experimenter and builder. They can be used to make compact equipment with excellent performance at low cost.

ham radio

fig. 6. 455-kHz i-f amplifier using MFC4010.





improving the fm repeater transmitter for amateur use

These simple modifications increase circuit Q and provide improved performance through lower receiver desensitization

J. Jay O'Brien, W6GDO, 6606 Fifth Street, Rio Linda, California 95673

A large number of amateur 144-MHz repeaters and remote-base stations make use of the Motorola PA-8491 60-watt high-band transmitter. This transmitter is ideally suited for repeater use but a few simple modifications will greatly improve its operation.

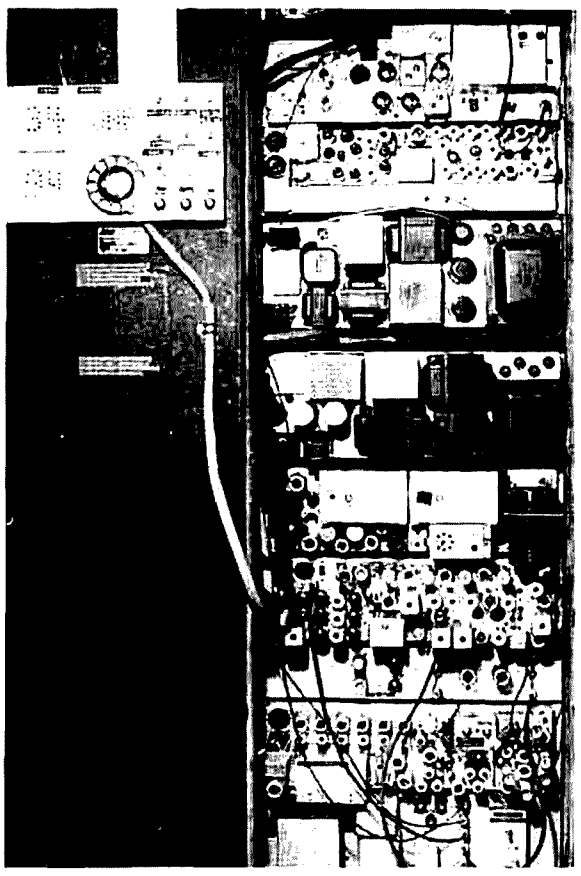
This transmitter, similar to others produced by Motorola and others, uses relatively low-Q tuned circuits. The low-Q circuitry permits moving the transmitter frequency a megahertz or more without retuning, and also permits replacement of oscillator and multiplier tubes with essentially no retuning required. In amateur repeater service these advantages, however, are far overshadowed by the problems generated by the broad spectrum of noise that the low-Q circuits permit the transmitter to emit.

The "white noise" is the cause of most receiver desensitization in amateur repeaters, especially in close-spaced systems that receive on 146.34 MHz and transmit on 146.76 or 146.94 MHz. The broad noise spectrum from the transmitter covers weak

signals on the receive frequency, causing the repeater receiver to be less sensitive with the repeater transmitter on than it is when the transmitter is off. This problem was covered very well by K5ZBA.[1]

This desensitization usually results in repeater "pumping." This occurs when the signal received at the repeater receiver is strong enough to actuate the repeater when the repeater transmitter is off, but is not strong enough to keep the repeater on when the repeater transmitter (and its white noise) is on. Under these conditions the repeater transmitter comes on, generates white noise that causes the repeater receiver to lose the incoming signal, and the repeater transmitter turns off. As soon as the transmitter is off, the receiver again detects the incoming signal and the cycle repeats itself.

fig. 1. Grass Valley repeater station, Wolf Mountain, California. This repeater covers most of the Sacramento Valley. The uhf and vhf equipment is shown here; decoder equipment is in another cabinet.



photos by WA6ESA

desensitization measurements

While preparing a repeater remote station package for mountaintop installation, desensitization tests were made under laboratory conditions. The 146.94 MHz transmitter was fed into a dummy load through a Bird Thruline wattmeter and adjusted for 60 watts output. A probe (such as the iso-tee[1]) was coupled to the dummy load, through a 10-dB 50-ohm attenuator to a milliwatt meter; the probe was adjusted for 60 mW indicated on the milliwattmeter (this represents a 30 dB loss from the transmitter to the milliwatt meter).

The connection from the 10-dB attenuator to the milliwatt meter was then disconnected and connected instead to the receiver input through a precision zero-to-50-dB 50-ohm attenuator. A signal generator was coupled through a 40-dB-loss probe into the line between the zero-to-50-dB attenuator and an unmodified Motorola Sensicon A receiver.

Receiver sensitivity was measured through the probe with the transmitter off; then the transmitter was keyed, and the zero-to-50-dB pad was adjusted to produce a loss in sensitivity of the receiver. This loss in sensitivity (desensitization) was measured as the number of dB of signal power increase required to produce the same amount of quieting (20 dB was used for these tests) as when the transmitter was off.

This measurement was performed with various amounts of transmitter power fed through the zero-to-50-dB attenuator into the receiver, thus simulating different amounts of antenna coupling loss; this loss is equal to the zero-to-50-dB attenuator setting plus the 30-dB probe loss.

It was initially determined that before we modified the repeater transmitter 79 dB of coupling loss produced 3 dB of receiver desensitization. This is equivalent to more than 60 feet of vertical separation between dipole antennas for receive and transmit. After modifying the transmitter, 41 dB of coupling loss produced 3 dB of desensitization; this can be obtained with 15 feet vertical separation between di-

poles, and represents 38 dB improvement.

Note that for these tests coupling from transmitter to receiver must be only thru the zero-to-50-dB attenuator. Direct radiation coupling must be incidental. This can be accomplished by separating the transmitter from the receiver in the rack, and placing other equipment in between. Shielding was not found to be necessary.

Transmitter modification consists of raising the circuit Q of the intermediate tuned circuits. This is done by adding capacitance across the tuned circuits in question, and in one case by reducing the coupling between tuned circuits. It is possible to obtain a couple of dB of additional benefit by critical slight detuning of the transmitter tuned circuits, however, this added improvement is not generally worth the added difficulty of adjustment.

transmitter modifications

The necessary transmitter modifications are quite simple, and are shown in the accompanying photographs.

A schematic diagram of the Motorola PA-8491 transmitter (less the filament circuit) is included on page 54 of the "FM Schematic Digest"; $3.95 from Two-Way Radio Engineers, Inc., Dept. HR, 1100 Tremont Street, Boston, Massachusetts 02120

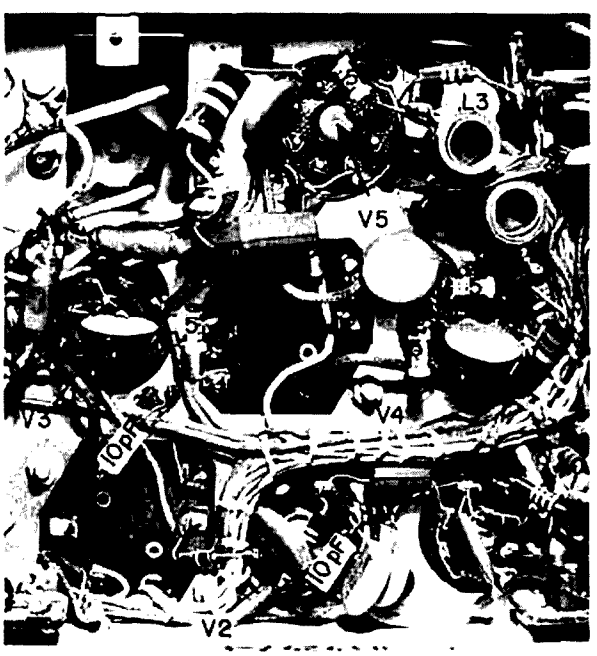
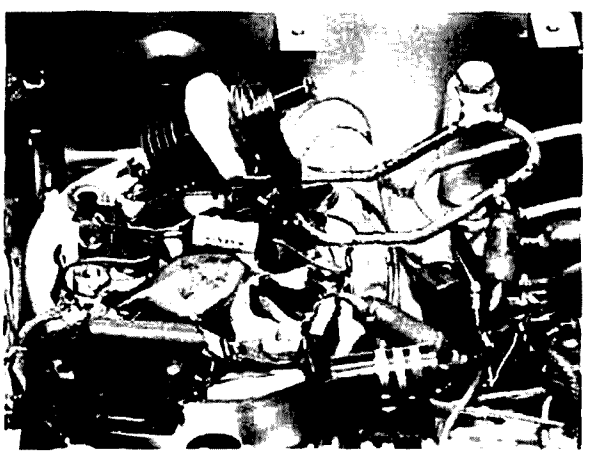


fig. 2. Bottom view of chassis, showing the added components.

1. Install two high quality 10-pF tubular ceramic capacitors across transformer T1: one capacitor from the plate of V2 to ground, the other from the grid of V3 to ground. These new capacitors are shown in **fig. 2.**

2. Install high quality 5-pF tubular ceramic capacitors across T2, L3 and L4 as follows: from the plate of V3 to ground, from the grid of V4 to ground, from the plate of V4 to ground, and from the grid of V5 to ground.

fig. 3. This photo shows the 8-pF air variable which is installed across the grids of the final amplifier.



3. Install a 1-to-8-pF variable capacitor across the grids of the final amplifier. The E. F. Johnson 160-104 capacitor shown in **fig. 3** does an excellent job but any similar air variable should work as well. This modification increases grid drive, improves efficiency, and raises the power output of the 829B final. An insulated tool must be used for adjusting this capacitor.

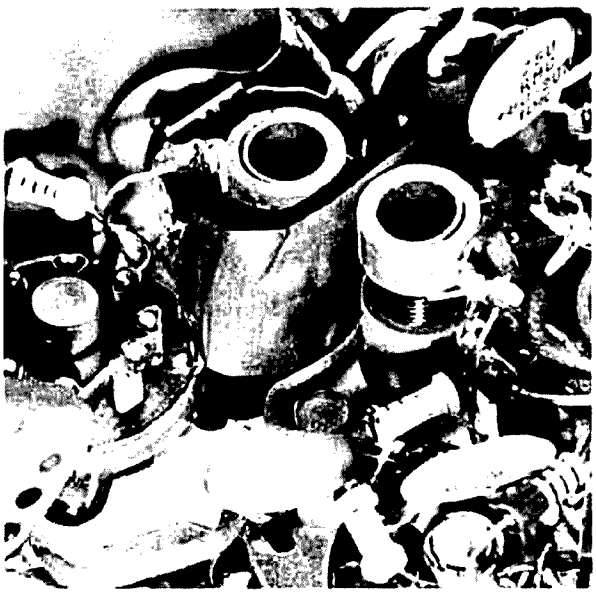
4. The coils at the plate of V4 and the grid of V5 (L3 and L4) are overcoupled, thus lowering the effective Q of this interstage network. Decouple these coils by mounting a small copper tab between them (**fig. 4**). Adjust the size of the tab so that after retuning, the driver grid drive is slightly less than before; an indicated drop of one microampere as measured at test position 5 should be about optimum. You will notice that the coils tune sharper

as you decouple them; this is an indication of increased circuit Q.

results

At our installation, we measure 64 dB of coupling loss between two three-element Yagis spaced 18 feet vertically on our tower. This is an acceptable figure for the modified transmitter. The top antenna is used for transmit on 146.94, and the bottom antenna for receive on 146.34 MHz. A 146.94 MHz receiver is also connected through the transmitter antenna relay to

fig. 4. The decoupling tab between coils L3 and L4 increases circuit Q.



the upper antenna. Both 146.94 and 146.34 are retransmitted as received to the control stations on the 440 MHz "down channel". No cavities are used in either the transmitter or receiver feedlines. With a weak signal (15 dB of quieting—about 0.3 microvolt) being received on 146.34, no change in limiter current or in quieting is noted by switching the 146.94 MHz transmitter on and off.

One note of caution, however; the swr must be low on both antenna feedlines. When the swr is high, the feedlines become part of the antenna, and since the feedlines usually run parallel to each other, the coupling loss between antennas is lowered.

measuring coupling loss

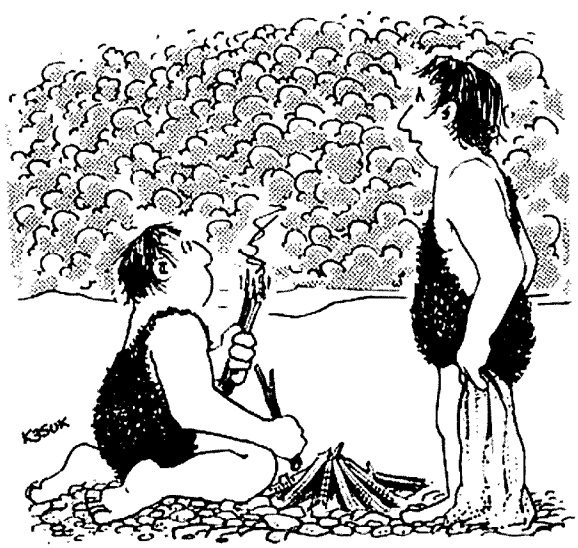
It is a good idea to measure the coupling loss between antennas to determine if it is adequate. The coupling loss between the two antennas may be measured by first connecting the signal generator (on 146.34 MHz) through a 6- to 10-dB attenuator to the 146.94-MHz antenna. Connect the 146.34-MHz antenna through a 6- to 10-dB attenuator to the 146.34-MHz receiver. Adjust the signal generator for a reference limiter current reading in the receiver. Now disconnect the attenuators from the antennas, and connect the attenuators to each other. The signal generator is then connected through two attenuators to the receiver. Reduce the generator attenuator loss until the reference limiter reading is again obtained. The difference in attenuator settings on the signal generator is the coupling loss.

reference

1. J. A. Murphy, K5ZBA, "Receiving System Degradation in FM Repeaters," *ham radio*, May, 1968, p. 36.

ham radio

a look at those days before spark... bs.



"Either the skip is open to California or the whole damn country's on fire!"

construction of high-frequency diversity antennas

Complete details on building new antenna designs described previously in ham radio magazine

George A. H. Bonadio, W2WLR, 373 East Avenue, Watertown, New York 13601

Four antennas of novel design and unusual characteristics have been described in **ham radio.**[1] Performance data and a brief description of physical arrangements and switching circuits were given. This article presents construction details on these antennas and their related components. Some observations are also given on how to get the best performance from the designs. My recommendations on feeder and element balance, as well as tuner and switching details, should be followed closely if you wish to duplicate the performance I've obtained.

Before deciding whether to build one of these antennas, it might be helpful to consider some of their important aspects. Initial cost is fairly high, but once installed and tuned, maintenance is no problem. My space-dimensional antenna has been performing well for several seasons with no attention.

Another thing to consider is that while these are not beam antennas they raise DX very well. An additional redeeming feature that justifies the initial investment is their excellent response to short-period selective fading. This is accomplished by a switching system that permits instantaneous pattern changes.

bonadio designs

The four antenna designs are the square diagonal, **fig. 1;** the box diagonal, **fig. 2;** the cube diagonal, **fig. 3,** which performs similarly to the box diagonal; and the space-dimensional antenna, **fig. 4.**

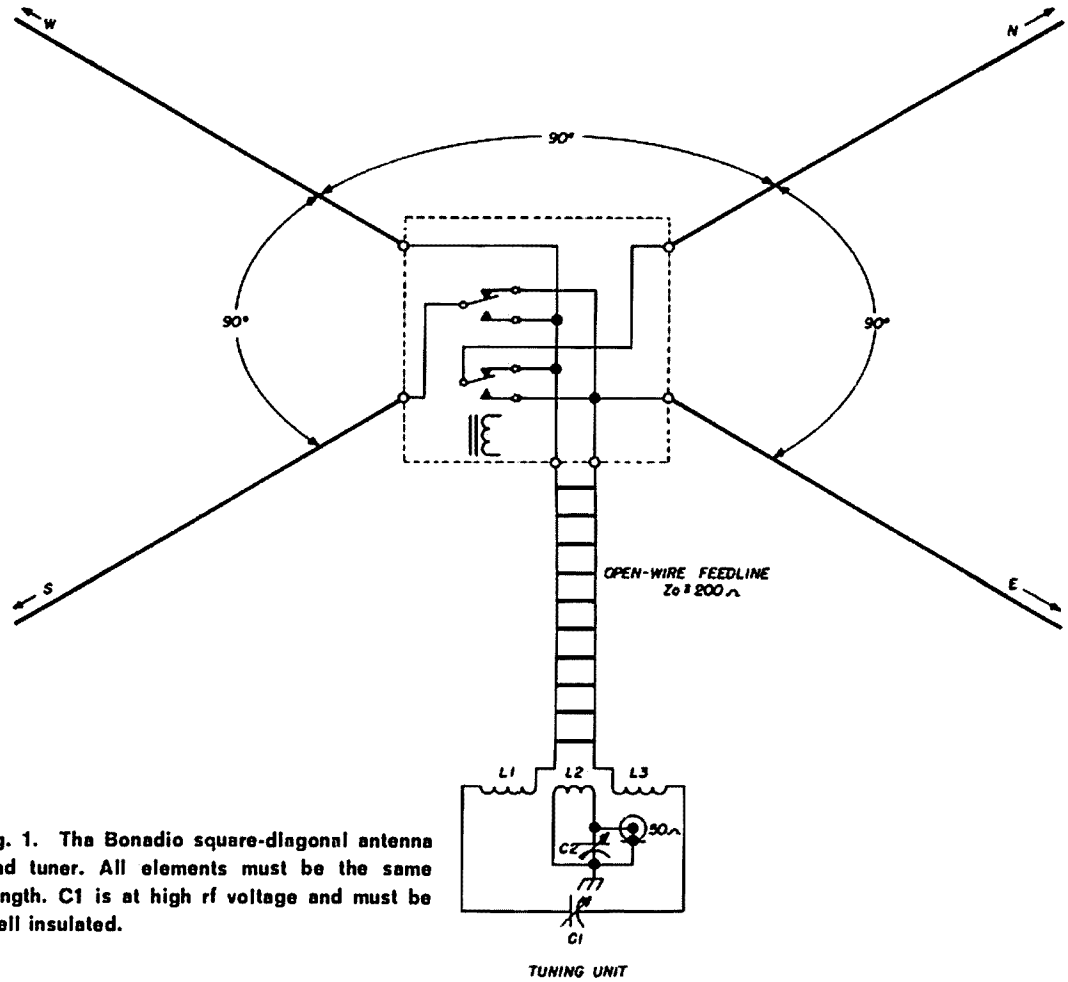


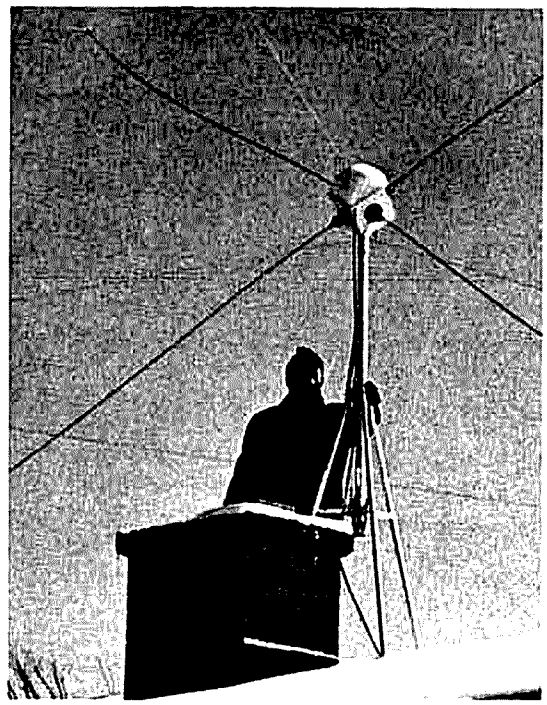
fig. 1. The Bonadio square-diagonal antenna and tuner. All elements must be the same length. C1 is at high rf voltage and must be well insulated.

characteristics

A glance at the diagrams shows many common parameters. The most important is the symmetrical relationship between element length, element spacing, support spacing, and angular element separation. The tuner system is the same for antennas. The coax link is series tuned above 10 MHz and parallel tuned for lower frequencies.

Compass directions are shown for antenna orientation. NU, NL represent north upper element, north lower element, etc. For the cube diagonal, elements must be spaced so that NU NL = NU EU = EU SU, etc. Angular separation of the cube diagonal elements is 70.5 degrees. While an excellent performer, the cube diagonal requires a rather complex switching system, and its nulls and general performance are eclipsed by the space-dimensional antenna. As with the other designs, symmetry and balance are a must in the space dimensional to obtain the deep nulls and optimum standing wave ratio over

Space dimensional antenna used in tests. It uses mobile mounts for elements, which are 9-foot whips whose ends are secured with 60-pound nylon fish line.



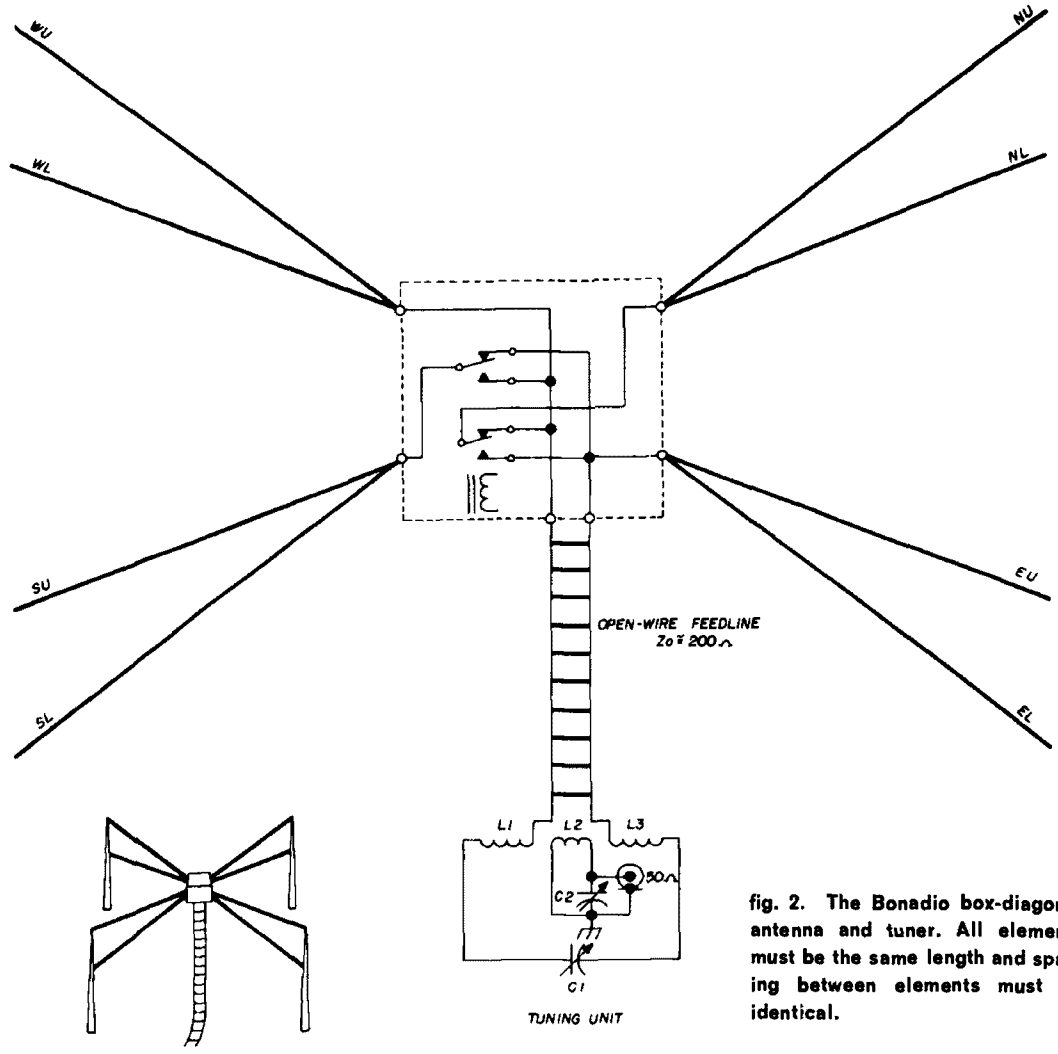


fig. 2. The Bonadio box-diagonal antenna and tuner. All elements must be the same length and spacing between elements must be identical.

its bandwidth. All angles for the space dimensional are 90 and 180 degrees, and all elements are the same length.

Another parameter common to Bonadio antennas is feed-point impedance. It is close to 200 ohms for very short wavelengths, with a small reactance component. The impedance varies in a cyclic fashion as wavelength increases. This is discussed in the description of antenna matching and transmission line design.

obtaining comparative data

To analyze the characteristics of these designs, you should have a control or comparison antenna. The best comparison antenna is the one you've been using. Don't worry about interaction between systems. There will be enough difference in reports between systems to provide valid data. The idea is to obtain as much quantitative data as possible over a given time period, because the larger the sample data size, the more realistic will be the final analysis. You won't need a computer to influence your opinion, nor will you have to rely on qualitative statements of others who might be using the same system. A large quantity of signal reports will convince you.

making a choice

Consider your antenna space. It can be anything from your bedroom ceiling to the back 40 acres. Study the types of supports needed for each of the four antennas. If you plan on three tall and three short flagpoles, for example, you're aiming at a terrific antenna, but a terrific cost for supports. Perhaps by relocating and redesigning your installation, you can borrow some buildings or other towers and trees. If you have the space, guyed towers are relatively inexpen-

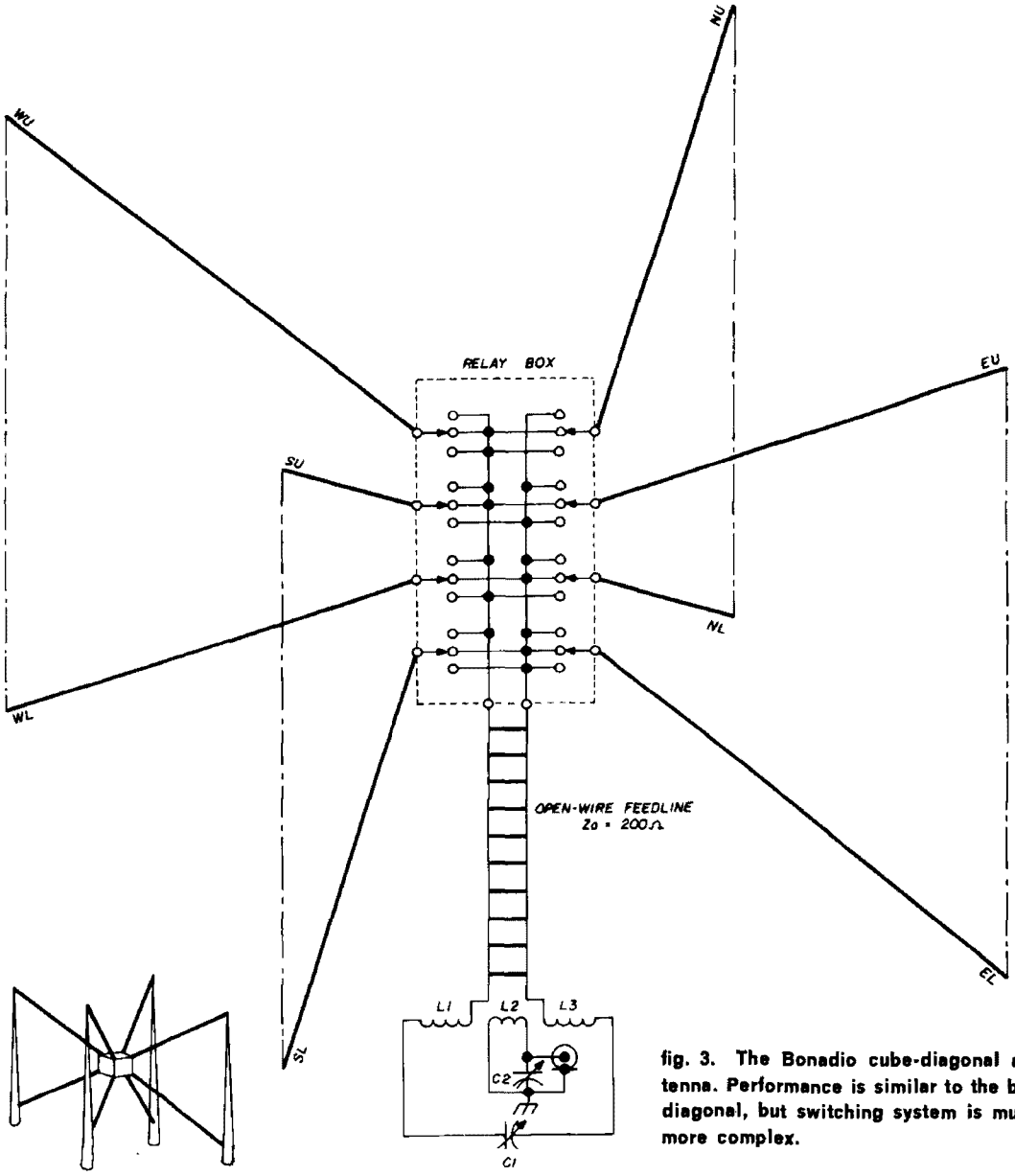


fig. 3. The Bonadio cube-diagonal antenna. Performance is similar to the box diagonal, but switching system is much more complex.

sive. Guy wires should be broken every twelve feet with strain insulators to avoid pattern discontinuities. Self-supporting towers are nice, but they cost a lot and require concrete bases. Flagpoles bear investigating.*

some definitions

Before starting construction, it's necessary to calculate the wavelength representing the lowest frequency at which an antenna of a given size will perform at full efficiency. The natural wavelength of an antenna will be termed λ_n. A wavelength that corresponds to λ_n plus 50 percent is called λ_L—the longest available wavelength at top efficiency. This is a parameter on which all Bonadio antennas are based and is extremely important. Here's how it is determined.

The λ_L of a square diagonal antenna is the distance around the perimeter of the square (which is its λ_n) plus 50 percent **(fig. 5)**.

The λ_L for either the box diagonal or cube diagonal is the distance around three sides of the top of the box or cube, then down both the two attached sides of the box or cube, then across the bottom connecting edge (this is its λ_n) plus 50 percent **(fig. 6)**.

The space dimensional antenna's λ_L is equal to the total length of its six elements;

* John E. Lingo & Son, "B" Division, 28th Street and Buren Avenue, Camden 5, New Jersey.

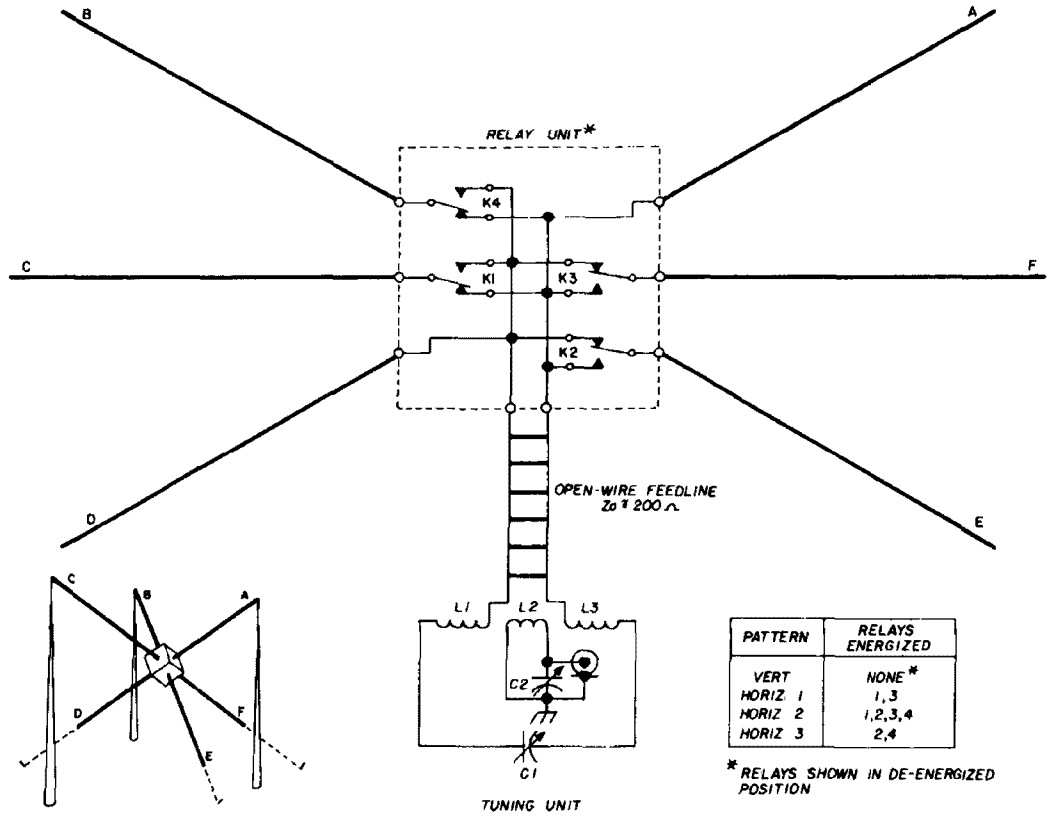


fig. 4. The space-dimensional antenna. Geometry and balance requirements are the same as for the other antennas, but performance is better. Three horizontal radiation patterns from this antenna are shown in fig. 11.

its λ_n is equal to the total length of any four of the six elements. If any of these antennas are driven with waves which are longer than its λ_L, it will still operate, but with less efficiency, about −6 dB per octave.

size losses

To determine the falloff of signal strength with antenna size, I compared a box diagonal with a space dimensional. The box diagonal antenna used 60-foot elements of number 8 aluminum wire, and the space dimensional had 10-foot elements and number 4 copper-wire feeders. These antennas, which were instantly switchable, were at a conveniently low height of 22 feet above ground at their centers.

On 20 meters, the antennas performed essentially the same (**fig. 7**). The space dimensional, with an λ_n of 23 MHz and λ_L of 15 MHz, was better on 15 meters and 10 meters. However, as expected, it fell off on 40, 80 and 160. On 160 meters, some of this loss was due to an outboard loading coil (number 16 silver-plated wire) that ran very warm. The larger antenna being used as a standard also fell off, since it was operating beyond λ_L. Also, as expected, the larger antenna exhibited a slightly smaller range of fading depth.

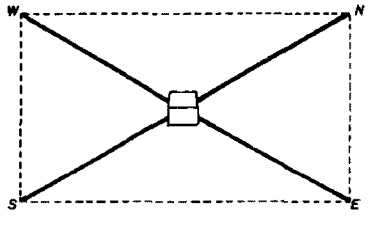


fig. 5. Resonant wavelength of square-diagonal antenna equals NE + ES + SW + NW.

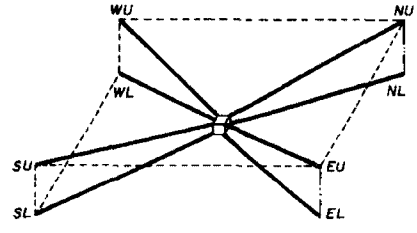


fig. 6. Resonant wavelength of box-diagonal antenna equals NU EU + EU SU + SU SL + SL WL + WL WU + WU NU.

Lay out your four, six, or eight elements side-by-side. Cut them all the same length. Connect them to their insulators using the same ties and the same length of wire. This balance is most important. An unbalance here will change your swr during pattern switching, because your elements will present different impedances to the feeders. On a box diagonal, the angular separation of elements must be identical. An error of 5 degrees will unbalance the system.

er than 66 feet, your near-in hop on 10 meters begins to suffer. A little higher, and near-in hop is reduced on 15 meters; when you go still higher, it is reduced on 20 meters. When your average, or center, elevation is below 60 feet, every few feet means a greater loss than the last few feet. You can't afford to operate below 16 feet for sheer power waste. Neither can you afford to spend money for a little better DX by increasing height to 100 feet.

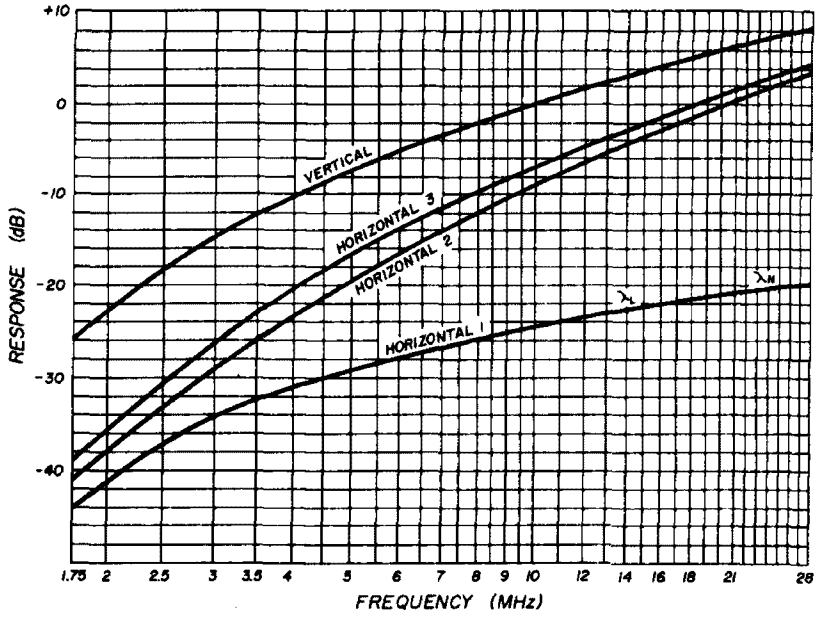


fig. 7. Comparison of two antennas; reference antenna at zero dB. Spread varies from approximately equal performance on 20 meters to about 40 dB difference on 160 meters.

wire sizes

All calculations I've seen on wire skin losses show that antenna elements or feeders smaller than pipe sizes will cause significant losses. There is, however, a diminishing return. I compromised on number 8 aluminum wire for elements and feeders on my larger antennas. On the small space dimensional (used in the tests above) I deliberately extended the number 4 wire feeders to 100 feet to eliminate loading coils through 80 meters. However, number 12 aluminum wire (used for tv grounding) should be satisfactory.

elevation

The advantages of elevation are greater for DX than for near-in hop. When you go high-

The feeder wire size should be consistent with element size. Commercial 200-ohm line is available from Federal Wire Company. However, I can't recommend it for these antennas except for low power.

I used a pair of 93-ohm RG-62/U coax cables, taped together, to feed relays without chokes, **fig. 8.** I found it too wasteful for all-band use; that is, for longer than 1/5 λ_L. The impedance is 186 ohms.

relays

There are no ideal switching relays for these antennas. You must use your ingenuity. I used up several Potter and Brumfield KT antenna switching relays whose insulation broke down from excess voltage at wavelengths beyond λ_L. My 140 watts of a-m de-

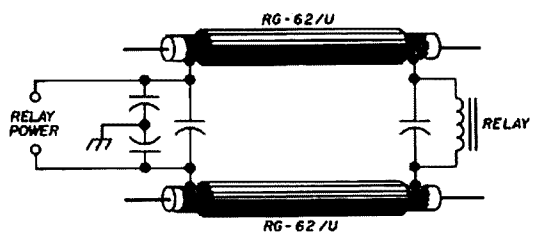
manded a larger relay, the Kurman 252C. This relay operated satisfactorily through rain and snow with the weather hood lost for two winters.

The Jennings Radio Company voluntarily furnished me with two engineering samples of their vacuum relays. These relays required no service after two more winters, but the amateur net price is over $200 per pair. They can switch any amateur power safely, even while transmitting.

The circuits I finally used are shown in **figs. 1, 2, 3** and **4.** For the cube diagonal, I have a Ledex number 250-124-256 relay that can handle only enough power for testing purposes.

The switching system for the space dimensional antenna, **fig. 4,** puts the coils of relays 1, 3 and 2, 4 in parallel. The upper antenna elements, ABC, are opposite the lower elements, DEF, in that order. A three-wire line is used for relay control. The rf chokes

fig. 8. Combination 186-ohm feeders and relay power line. This system is only suitable for very low power.



shown in the photo of the open control box are in the power leads to the relays. Plastic sleeving is used over leads between chokes and coils. These chokes prevent rf from arcing across relay coil leads. The chokes are 10 mH each. Note particularly the physical arrangement of the relays.

relay power

It is possible to power the relays with 120 Vac, but this generates noise. I've used it through my feeders; however, on dewy mornings there was an S3 power-leak noise in the receiver when the relays were energized. Conversion to 120 Vdc corrected this. A word of caution: don't use powdered iron core rf chokes. One of mine became red hot with only 100 watts input. Bypass the relay coil with an 0.01 μF ceramic capacitor to be safe. The open-wire switching system is shown in **fig. 9.** Accidental short circuits can damage a choke, because the power supply fuse won't blow. The circuit shown isn't practical for short feeders.

For my compact space-dimensional antenna, I used a three-wire cable to switch the four antenna patterns shown in **fig. 7:** horizontal 1, horizontal 2, horizontal 3 and vertical. Relay switching logic for the patterns is shown in **fig. 4.**

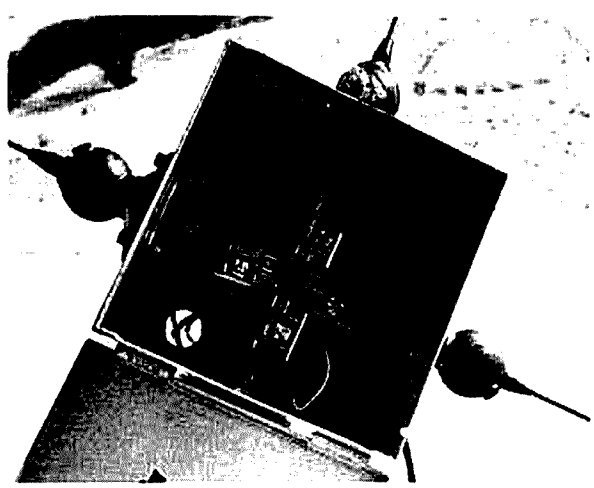
feeder construction

The open-wire 200-ohm line uses Reynolds Metal Company type 2B tv insulators, **fig. 10.** Protected line requires insulators every two feet; swaying line needs them every six inches minimum. Thread the insulators half way down the line from each end, using mineral oil lubricant if necessary; then space them appropriately. If the line is anchored every four feet, insulator spacing can be tripled. Losses are so low that a thousand feet can be used.

matching system

The feed point of these antennas, on the very short wavelengths, looks close to 200 ohms and has very little reactance. However, as the wavelength increases to λ_n the impedance may reach 20 ohms after passing points of over 1,000 ohms. On wavelengths

Control box for space dimensional antenna. The rf chokes prevent rf from arcing to relay coils and leads. Three-wire power cable enters box at lower right.



longer than λ_L, the impedance may fall to less than one ohm, and the capacitive reactance may reach hundreds of ohms. Meanwhile, the standing wave ratio can exceed 200:1 on the balanced feeders, while maintaining 1:1 on the unbalanced coax.

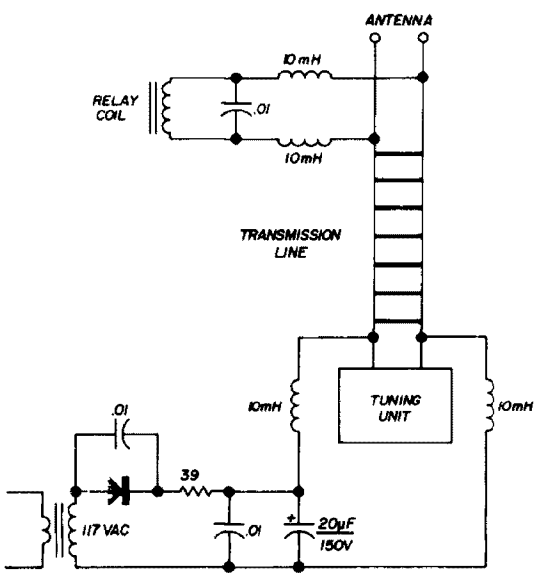
An open-wire line of 160 to 250 ohms impedance is optimum. The modest mismatches of the transmission line will probably produce more feeder-created reactance at the tuner feed point than the antenna displays alone, at anything shorter than λ_L. These broadbanded, low-Q low-reactance combinations are much easier to manage than the common center-fed open-wire feeder antenna systems.

tuners

To deliver all of your available power into feeders, you should use the tuner circuits I have shown, **figs. 1** and **2.** With these tuners you can always tune and load through a standing wave ratio of 1:1. With other tuners you can hit a standing wave ratio of 1:1 quite often and keep a cool tuner, but on some frequencies your tuner will not deliver much of the power to your antennas.

Don't use the tuner of Handbook X, or Handbook Y, or manufacturer Z, and then condemn the antenna as "not especially good."

fig. 9. Providing relay power through the open-wire transmission line. Iron-core chokes should be avoided; power is limited to 250 watts.



tuner values

Table 1 gives optimum tuner values for a 200-ohm resistive load. If λ_L is much greater than optimum for the bands shown, the values will still be good for the larger antennas. However, you may have to change the values somewhat with the smaller versions as you approach λ_L. At wavelengths much longer than λ_L, variations from values for L1, L3 and C1 may be large. The link values should be as shown. They are designed for 52-ohm coaxial cable.

A surprise benefit of these tuners is the small air-gap sizes of the capacitors. The

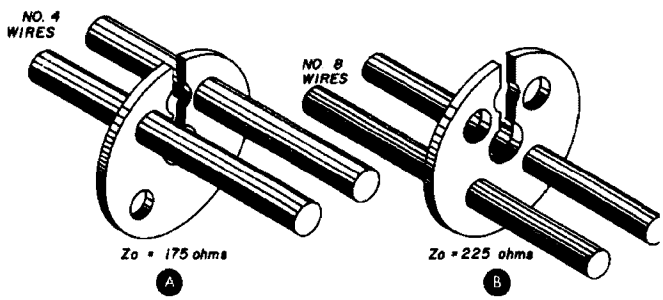


fig. 10. 200-ohm open-wire transmission line using Reynolds Metal Company's type 2B tv line insulators. If suspended in space, a spreader is required every six inches for number 8 wire; every 12 inches for number 4 wire.

parallel- and series-tuned links must safely handle rf peaks of 350 and 1000 volts, respectively, for full legal power.

Antenna tuning capacitors with a 3000-volt rf peak rating will safely handle full legal power. However, on some bands this requirement will increase to 8000 volts for a kilowatt.

The tuner-bandswitching apparatus shown in the photo has given me no trouble over the past years. I used a speaker box for an enclosure. An attempt has been made to keep the impedance constant by using braid paths equal to the center path of the coax. The coils are made of silver-plated copper wire. Intercoupling is insignificant, because coil resonant frequencies are well separated.

You'll find that a gang switch is more desirable than conventional tuner circuits. You'll have optimum tuning, coupling, and loading control at the flick of a switch. As the tuner is broader than a tank circuit, you

can listen across an entire band without retuning. Because the system is double tuned, images are down 30 to 60 dB; generally to extinction. Also, you can load your transmitter quickly anywhere near the last operating frequency on each band.

loading

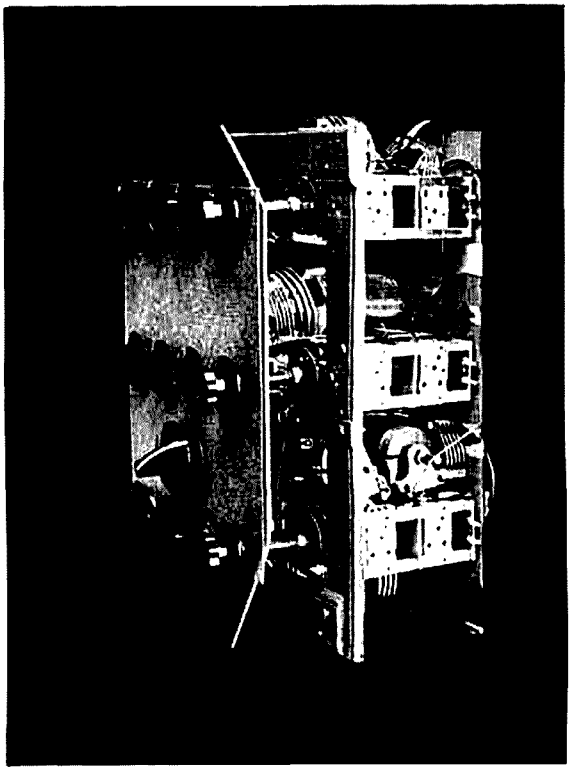
If a tuner resonates but refuses to load at maximum coupling, there are several cures. You can reduce the clearance for L2, between L1 and L3; increase the number of turns equally on L1 and L3, and reduce the capacitance of C1 accordingly; or reduce the lengths of L1 and L3 by using the next smaller wire size to obtain slightly more inductance.

Proper coupling occurs when the swr passes through 1:1 with the link between one-half and fully in. If the swr goes through 1:1 before the link is one-half in, then L1 and L3 have too many turns, and C1 may arc over.

You can wind the links of number 8 solid copper wire, or 1/8-inch tubing (both silver plated, if possible), over a two-inch-diameter bottle. Space the turns about one-half conductor diameter. The links swing between the antenna coils, which are made of number 14, 16 and 18 wire.[2]

The fixed padder-type capacitors are Arco-El Menco series 30.* In parallel-tuned circuits these will handle 140 watts of a-m power. With series tuning, however, this amount of power can cause corona discharge. You can stack these capacitors, using two layers of mica in place of one. For power up to a kilowatt, double the layers of mica. This lowers the capacity, which requires more plates or additional capacitors. Caution: don't use ceramic or moulded micas; they've been known to fail.

* Allied Electronics 43A7093 through 43A7106 for 2 through 15 plates (130 through 3055 pF).



The tunar-bandswitching apparatus. All coils are silver-plated. Tuner mounts in a speaker enclosure.

link tuning

Each link is for one band only. At any frequency in the band, the link will present an insignificant reactance to the mutual coupling complex. Operating Q of the links is about three.

table 1. Component values for antenna tuners shown in fig. 1, 2, 3 and 4. L2-C2 is parallel tuned on 160, 80 and 40 meters, and series tuned on 10, 15 and 20.

amateur band	L1 (μH)	C1 (pF)	L3 (μH)	L2 (μH)	C2 (pF)
160	80	32 - 160	80	1.6	4700
80	40	16 - 80	40	0.8	2400
40	20	8 - 40	20	0.4	1200
20	10	4 - 20	10	1.6	68
15	7.5	3 - 15	7.5	1.2	47
10	5	2 - 10	5	0.8	33
6	3	1.2 - 6	3	0.5	22
2	1	0.4 - 2	1	0.16	6.8

Grid dip the links with the coil and capacitor in parallel. For final trimming, the link will indicate a dip when placed at least four inches from the grid dip meter's coil. Thereafter, the series links are opened on one side, and the coax is fed between them. The padder capacitors are never adjusted again.

To check for balanced operation, tune the receiver to a strong signal on each band. Short circuit the feeders while watching your s-meter. Signals should drop 30 to 60 dB. A drop of less than 20 dB indicates poor balance, which should be investigated.

Full-sized Bonadio antennas have low Q. When changing frequency you'll have to retune your final amplifier three times as often as your antenna tuner. However, if you use an antenna whose wavelength is longer than its λ_L you'll have to retune the antenna tuner more often.

patterns

The horizontal radiation patterns for the four- and eight-element systems are shown in **fig. 11.** For many of the shorter skip paths, the nulls will be —10 dB or less when signals are coming in at high angles. Occasional instantaneous fading differences will exceed —30 dB.

For the space dimensional, the patterns for H1, H2 and H3 are all 60-degree overlapping figure 8's. Line noise is most pronounced with the vertical pattern. Often one H pattern will decrease line noise by 20 or 30 dB. One day the BBC 21.47 MHz echo garble disappeared with an H-pattern null to the north. Later the same day, the BBC signal's flutter smoothed with an H-pattern null to the south of their normal path.

fading

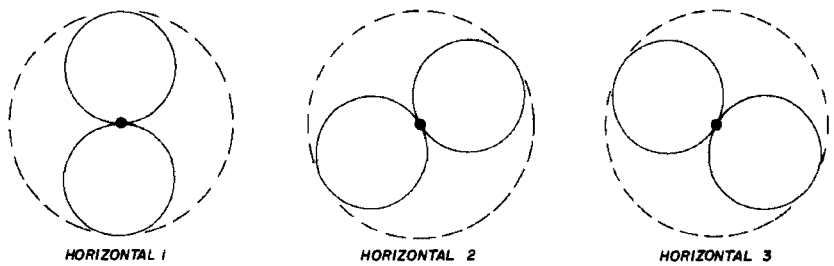
The greater the DX, the greater will be the pattern change differential. On near-in hop, differences are mostly washed out by fading of the received signal, even though your contact tells you your signal has almost no fade.

You'll find that the control of fading signals is a matter of switching your patterns. If the received signal fades on pattern H1, immediately switch to pattern H2. When the H2 signal starts to fade, go back to H1, H3 or V. This is diversity reception on a one-antenna system.

You'll receive fewer reports of fading than other stations. This will be most apparent during round-table contacts, because the difference will be more conspicuous. If the received signal is also from a Bonadio antenna, selective fading will be the least of your problems. You have to experience it to believe it.

As you might expect, for DX the vertical pattern usually is equal to or better than the best horizontal pattern. However, in receiving, certain interferences may be reduced by one of the H pattern nulls, so you may find an H pattern preferable. The V pattern is helpful for quickly finding which area of DX is coming through best.

fig. 11. Horizontal radiation patterns for the space-dimensional antenna. Element switching for each of these patterns is shown in fig. 4.



The Bonadio square-diagonal antenna is protected by U. S. Patent 3,274,606. Patents are pending on the Bonadio box-diagonal and cube-diagonal antennas; the inventor has applied for a patent on the space-dimensional antenna. This protects the inventor against commercial infringement but does not prevent an amateur from building one for his own use. **Editor**

references

1. George A. H. Bonadio, "A Survey of High-Frequency Antennas," *ham radio*, April, 1969, p. 28.
2. Illumitronic Engineering Corp., 680 East Taylor Avenue, Sunnyvale, California. Section 1800, eem file, p. 6, 7.

ham radio

solid-state exciter for 432 MHz

Here's a solid-state exciter that converts 20 mW of two-meter drive to 22 watts on 32 MHz

Hank Cross, W1OOP, 11 Birds Hill Avenue, Needham, Massachusetts 02192

In twenty-one years of operation on 432 MHz I've had nine different vacuum-tube exciters, usually driving one or more 4X150A's. The latest of these, veteran of three winning mountain-top contest operations, is an excellent example; it uses a 6AN8, 5763, two 6360's and a 4X150. Heater power is 33 watts and it requires 100 watts dc on transmit.

Presently I'm using a solid-state exciter (**fig. 1**). If it were directly crystal-controlled it would use six or seven transistors plus a varactor. Dc input power is 50 watts on transmit and it has enough output to do the same job the tube exciter did. Including vfo and power supply, it takes 5¼ inches of rack space.

vhf power stage

Before getting into any further details, lets talk about the typical vhf power-amplifier stage shown in **fig. 2.** It uses a 2N3375 operating class-C at about 144 MHz. Drive power is about one watt and output is about six watts, with a 24-volt supply (at twelve volts the power gain and power output will be about half).

The 2N3375, the most common of the overlay transistors, can be thought of as 144 type 2N706's in parallel, all assembled on a chip 1/16-inch square, with the corresponding elements hooked together by a network of evaporated-aluminum bus bars in the usual integrated-circuit manner. The collector,

the body of the chip, is gold-soldered on top of a beryllium-oxide wafer which is soldered to the copper stud for cooling. With a coral-colored aluminum-oxide top insulator and gold-plated metal parts, it looks like a piece of costume jewelry.

The base is fed from a low-impedance source. There should be an effective short circuit for harmonics of the drive frequency so that the base current can flow in short pulses. Capacitor C3, which accomplishes this, reduces the dc drain without affecting power output; it's handy to make it variable. The input impedance is 10 ohms or so in series with a bit of lead inductance.

To get power out with less than 24 volts peak swing, we need a load around 35 ohms. Coil L3 resonates with the transistor output capacitance, which will be about twice the rated C_{ob} value under full power (remember that capacitance changes non linearly with collector voltage). The Q of this resonant circuit is around one, so the value of L3 is not very critical. Also, the collector voltage waveform will not be sinusoidal, but rather more like an unsymmetrical square wave. Harmonic rejection is provided by the series-resonant circuit L4-C4 while C5 is used to transform the 50-ohm load down to the needed 35 ohms. Capacitor C5 should have a reactance at mid-range of about 60 ohms and L4 around 200 ohms at the operating frequency for a Q of 5 or 6.

When things are working right the load line will be approximately as shown in **fig. 3.** If the transmission line breaks or shorts at exactly the right spot, the load could become a dead short (as viewed from the collector) and the chip will melt in the time it takes twenty or thirty watts to heat 0.00004 cubic inch of silicon four hundred degrees. A protective circuit to avoid this consequence is shown as part of **fig. 4.** It was suggested by K1BRO and is similar to one found in the RCA transistor handbook.[1]

The choke in the base circuit, L2, should be low-Q. I used a three-turn toroid wound on a ferrite bead, but several inches of fine wire wound around a 150-ohm resistor will also do the job (the resistor is in parallel with the choke, parasitic-suppressor style).

The feedthrough capacitor, electrolytic capacitor and 0.47-ohm wirewound resistor (or ferrite bead) are also important. The electrolytic should be solid-tantalum type, although ceramic or extended-foil paper capacitors as low as 0.05 μF would probably be adequate.

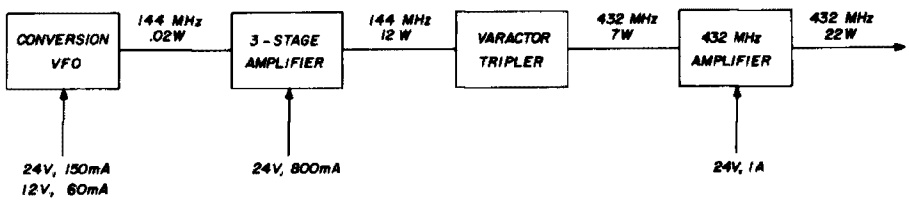


fig. 1. Block diagram of the solid-state 432-MHz exciter.

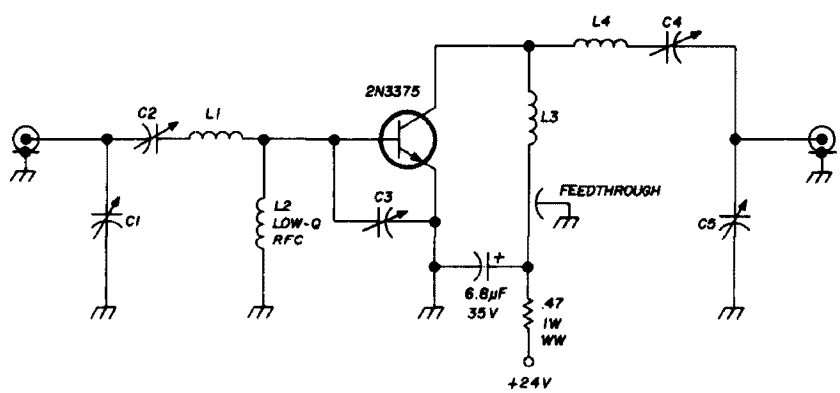
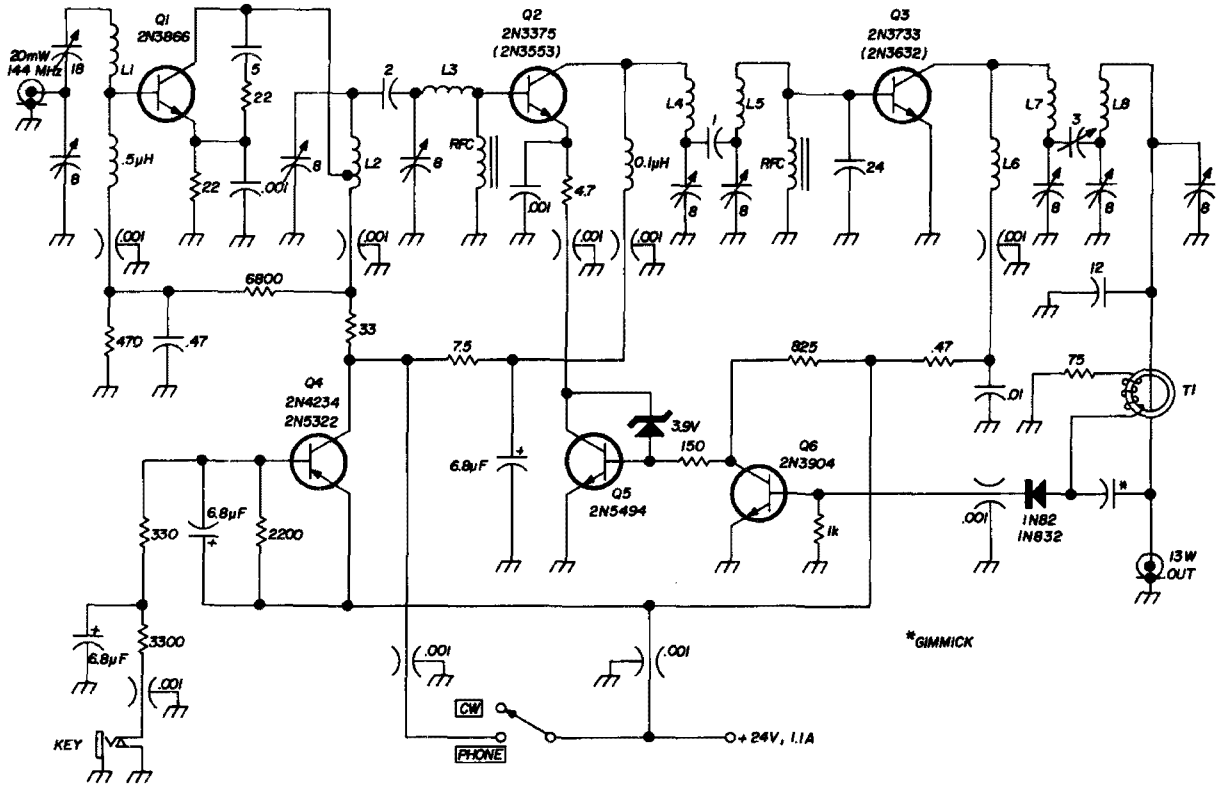


fig. 2. Typical class-C vhf power amplifier stage.



L1	**4 turns no. 20, 0.2" diameter**	**L7, L8**	**8 turns no. 24 enamelled, 0.2" diameter, 0.2" long, tunes to 144 MHz with 4.5 pF**
L2, L3	**5 turns no. 20 enamelled, 0.2" diameter; tunes to 144 MHz with 8 pF; L2 tapped at 2 turns from cold end**	**T1**	**Small *toroid* slipped over line to coax connector; 1 turn for pickup (Ferroxcube 213T060 in 4C4, 1Z2 or 1Z3 material, or Ferramic Q-3)**
L4, L5	**8 turns no. 24 enamelled, 0.2" diameter, 0.2" long; tunes to 144 MHz with 4.5 pF**	**RFC**	**3 turns no. 26 through a ferrite bead such as 3B7 shielding bead**
L6	**8 turns no. 24 enamelled on a 1/4-W 100k resistor**		

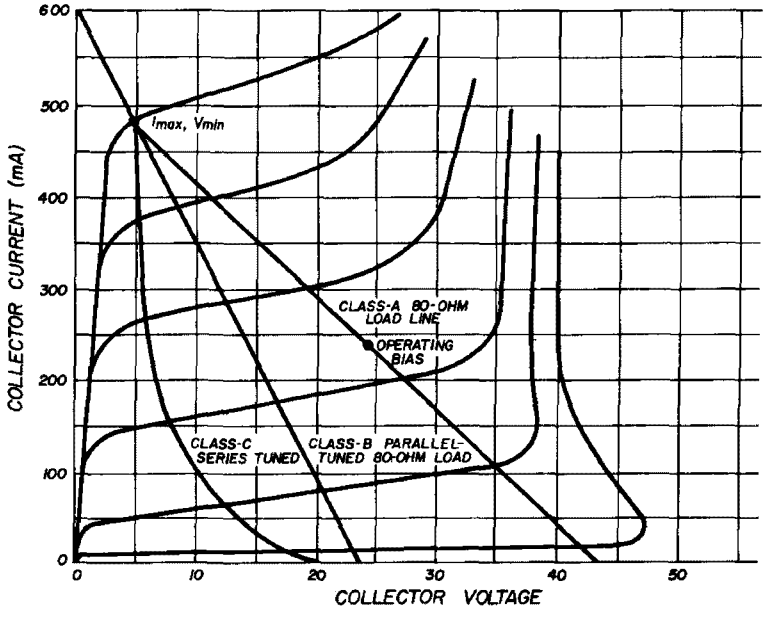
fig. 4. Three-stage 144-MHz amplifier provides 13 watts output with 20 milliwatts drive. The built-in vswr detector protects the final transistor stage against serious mismatches.

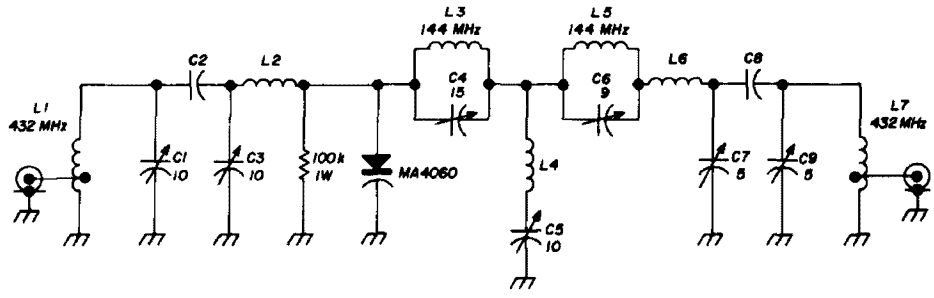
Emitter grounding is a major problem; the stripline package brings out two leads at 180 degrees. **Any** emitter inductance is **too much.** A sixteenth of an inch will give an ohm of inductive reactance at two meters, enough to seriously affect gain or possibly result in parasitics.

two-meter amplifier

The three-stage 144-MHz amplifier is shown in **fig. 4.** The first amplifier, Q1, is operated class AB to get more power gain. The 2N3866 is very "hot" and tended to have parasitic oscillations; there are several additions to the circuit which were put in to tame these parasitics. A 2N3375 was used for Q2 because it fit my layout better than the less costly 2N3553 (for salvaged ones, the price was the same); it provides about

fig. 3. Load lines for rf power amplifiers.

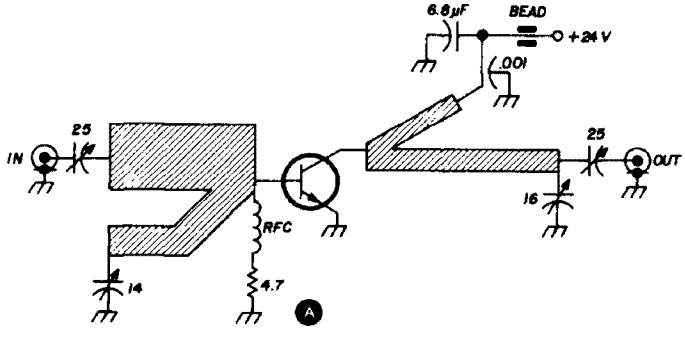


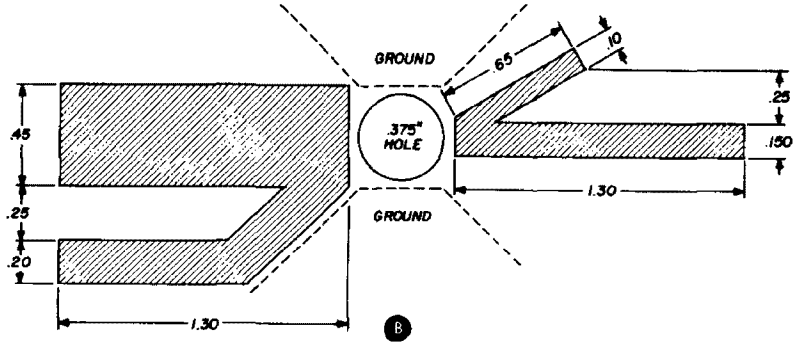


C1, C3, C5	10-pF miniature variable (Hammarlund MAC-10)	L2	7 turns no. 18, 3/8" diameter, 1/2" long
C2, C8	leads of no. 26 insulated wire, twisted together for 2 turns	L3	4 turns no. 22, 1/4" diameter, 5/16" long; tune cold to 144 MHz
C4	13-pF subminiature variable (E. F. Johnson 189-6)	L4	4 turns no. 18, 1/4" diameter, 3/16" long
C6	9-pF subminiature variable (E. F. Johnson 189-6)	L5	1-1/2 turns no. 22, 1/4" diameter; tune cold to 288 MHz
C7, C9	5-pF miniature variable (Hammarlund MAC-5)	L6	2 turns no. 20, 1/4" diameter, 1/8" long
L1	9 turns no. 18, 3/8" diameter, 1/2" long; tap at 2-1/2 turns	L7	3 turns no. 20, 1/4" diameter, 1/4" long; tap at 1-1/2 turns

fig. 5. Varactor circuit for tripling from 144 to 432 MHz.

fig. 6. The 22-watt 432-MHz power amplifier uses printed-circuit stripline inductors; material is 1/16-inch copper-clad Telite. The rfc in the base consists of 2 inches of wire airwound with an 8-32 screw as a form.





two watts output. The protective reflected-power circuit is hooked into its emitter.

Transistor Q3 was originally a 2N3375, but after blowing a couple I put in the larger 2N3733 at the same time I added the reflected-power control circuit. I guess either change would have been enough. The 2N3375, working hard, gave ten watts at slightly better efficiency than the 2N3733, which puts out twelve or thirteen watts. When tuning up at high specific power, turn the collector voltage on and bring the drive up gradually. Time constants can be set up to do that each time things are turned on.

When the reflected-power detector is working properly, the back power (as indicated on a good wattmeter) goes up to one watt and holds as the forward power is reduced to one watt with complete reflection. This means that a varactor multiplier may be tuned up with only an output power monitor (provided it's tuned to 432 and ignores 144-, 288- and 576-MHz energy). When tweaking five to seven adjustments it helps out if you don't have to watch the input matching at the same time. And if that last touch pops it out of tune, the drive transistor is protected. I first used the varactor tripler described in **QST**[2]. Then I built a

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similar one in a smaller box to fit the space available. The circuit is shown in **fig. 5.**

432 amplifier

The six or seven watts of 432-MHz output from the varactor is applied to a stripline amplifier **(fig. 6)** built according to information provided on the RCA-TA7344 data sheet[3]. The transistor I am using is a Fairchild developmental unit (TBLO504) rated for 25 watts output at 400 MHz with a 28-volt supply. I'm getting 22 watts at 432 MHz with a 24-volt supply. Other suitable types are the 2N5177 (TRW), MM1551 (Motorola), or TA7344, 2N5016 and now extinct 2N5017 (RCA). I am not using any mismatch protection in this stage at present.*

conclusion

So far, we have a carrier source. This might be used to drive a high-level mixer for sideband operation; it will work by itself on cw (a keyer is shown) or for a-m we can modulate the final 4X150. I don't think that a satisfactory linear sideband amplifier chain can be made to the ten-watt level with present transistors because of problems with linearity and thermal runaway, but devices to solve these problems will be on the market in a year or two.

For a-m operation on two meters, look at the data sheets for the RCA 40290, 40291 and 40292. The 40292 appears to be a 2N3632 at a reduced price, rated for a-m operation with a 14-volt supply. There are other types also designed for this sort of service. There are not any comparable 400-plus-MHz types that I am aware of, although I know that several manufacturers are making 225-400 MHz a-m transmitters for aircraft service. I'd probably learn a lot from an appropriate technical manual.

* Late note: I should have—I just blew the transistor!

references

1. "50-MHz, 40-Watt CW Transmitter with Load-Mismatch Protection," RCA Transistor Manual SC13, 1967, p. 504.
2. Henry H. Cross, W1OOP, Frequency Multiplication with Power Varactors at UHF," *QST*, October, 1962, p. 60.
3. RCA RF Power Transistor data sheet, developmental type TA7344, April, 1968.

ham radio

calculating received power in a radio communications link

A detailed analysis of just what happens to the hard-earned watts from your transmitter

Jim Ashe, W1EZT

The essential element in any radio communications link is the connection between some distant transmitting antenna and a receiving antenna. Perhaps you've never thought of a "connection" through space, but such a connection nevertheless exists.

The purpose of this piece is to analyze the radio frequency energy propagation problem and demonstrate how you can estimate the amount of power your antenna will pick up from a given transmitter, or to put it another way, the amount of input your receiver must have for effective reception.

approach to the problem

Such an estimate can be very helpful in choosing equipment to be purchased, planning a station design, or thinking about whether or not you want to try some tough new project. Maybe someday you might want to design an rf link between two sites. Or perhaps you're thinking about moonbounce or space communications.

As I reviewed the basics required for such work, I was very surprised to discover how simple it all is, **if approached properly.** Basically, received power questions are just another circuit problem, except we have to deal with power traveling through space as well as through wires and transmission lines. Space has only a few basic physical and geometric properties that are always the same, except near large stars. But that kind of space is likely to be far outside our experience for years to come. Let's now consider why received power and receiver performance questions sometimes seem difficult.

the rf spectrum

Historically, people haven't been working very long with radio signals that behave in a simple manner below the ionosphere. The most-used frequencies have been below

about 100 MHz. Radio signals don't behave at these frequencies in the straightforward manner of those in space. The frequencies we're interested in at present cover a wide range, amounting to five decades, with a lower limit of about 10 kHz. It's informative to consider the physical aspects of the radiating medium with which we must work at the frequencies of interest. The following summary presents the general characteristics of antennas in the entire rf range.

under 3 MHz

Antennas are electrically short and usually reactive. Elaborate grounding is required. Antenna input resistance is low, requiring a matching network; yet the antenna is physically large.

3 MHz to 30 MHz

Antennas are a large part of a wavelength in size, and grounding becomes optional. Input resistance is easily matched, and a structure that is electrically large is physically manageable. Some directivity can be achieved without mechanically huge, electrically phased systems.

30 MHz to 300 MHz

Antennas can be made in dimensions of a few wavelengths. Grounding is usually of no importance except for lightning protection. New kinds of antennas and wideband structures begin to appear. Very sharp directivity becomes possible along with high gain.

300 MHz up

Real variety appears in antenna designs. Some structures use optical techniques and resemble optical devices. Mirrors and lenses appear. Apparently strange means of carrying rf include the G-string and waveguide techniques. Antenna bandwidths may be measured in octaves, yet the antennas are very directive. At frequencies in the gigacycle region, complications begin to exist in such areas as the physical size of water droplets, which approach a wavelength in dimension.

The natural laws apply to all these antennas; however, there are so many differences it's somewhat surprising to recall that most involve practical problems of mechanical size and engineering. A waveguide, for example, would certainly work at 80 meters but would be huge, expensive, and impractical. A top-loaded, tenth-wavelength radiator with several downleads appears equally improbable at vhf.

Although our one field of antenna engineering has the appearance of several, I have chosen wavelength (or frequency) as the only variable in my analysis of the rf energy transit problem.

propagation modes

The communications engineering field is further complicated by the four basic classes of rf propagation. We have surface and space waves as pure types, ground and sky waves as the varieties we actually find at low frequencies, and a catch-all department for the multiple and once-unsuspected vhf modes discovered since 1940. Let's examine the characteristics of the rf propagation modes.

surface wave

Propagation of the radio signal is along the ground, suggestive of a one-sided waveguide. Propagation efficiency depends, in part, on ground characteristics, with sea water being least lossy.

space wave

Propagation is in straight lines through space—the simplest possible condition. Propagation through the atmosphere sometimes closely resembles space-wave propagation.

ground wave

Surface and space wave combine as they propagate at low frequencies. A vertical radiator is preferred with the bottom end well grounded.

sky wave

This is a low-frequency space wave as observed under conditions in the atmosphere. It must be low enough in frequency for reflection from the ionosphere. Under good conditions it may reflect from the ionosphere and ground several times before weakening to inaudibility.

vhf modes

If there were no atmosphere, vhf signals could be heard beyond the horizon only by diffraction scattering. This will be the case for explorers on the moon. In air various effects include scatter, TEM, auroral reflection, inversions, duct propagation and sporadic-E.

Suppose we have a long-distance radio link between New York and, say, London, Florida or Australia. A few of the factors that must be considered include:

1. Transmitting and receiving antenna properties, including grounding.
2. Path orientation in earth's magnetic field.
3. Daylight or darkness along path.
4. Season.
5. Solar activity.
6. Immediate atmospheric ionization.

Any experienced radio amateur could add entries and detail enough to fill the page, and a professional communications engineer could probably write a book on the subject.

If this description doesn't make estimating received signals a black art for super computers, here are still other factors: antenna orientation relative to the incoming signal; the relative phases of signals arriving by alternate routes; whether signals are polarized linearly or circularly; and their orientation or sense of rotation.

Finally, there is the competition of the received signal with external and receiver noise. Your receiver may provide excellent output from a weak signal on the bench, but connected to the antenna it also brings in 60-Hz harmonics from the millions of noise machines in cities, atmospheric noise at the lower shortwave frequencies, other amateur stations, and often strong interference from improperly tuned commercial broadcast, propaganda and jamming transmitters. At very high frequencies, the incoming signal must compete with thermal and atomic noise from the universe.

Now that I have made this sound like a practically impossible problem, I'll devote the rest of the article to an explanation of how satisfactory solutions to the radio-link problem may be worked out with fifth-grade arithmetic. Some tenth-grade algebra will be of assistance in the more difficult cases, and the key ideas are of roughly the same level of difficulty. Our approach will depend upon key points.

The first is: **accuracy** is not required. Twisting a knob or two on your receiver will throw in gain or loss fudge factors of a thousand or so per knob, and your receiver's agc control is comparably effective. A signal-strength estimate off by 10 or 20 dB will still be very useful and tremendously better than a guess or no estimate at all.

Secondly, if you do come out wrong, your **notes** will help set things right with least effort, since records and experience combined will frequently inform you just how much improvement is needed and where it can be made most easily.

Thirdly, you probably don't appreciate the really tremendous **inefficiencies** your system can have and still work. Normal communications systems commonly deliver larger amounts of signal power than are really needed. This brute-strength advantage is almost unavoidable under normal operating conditions, unless you're working with very, very low power. Take a note from the low-power amateurs, who cover hundreds of miles on milliwatts. Modern required-power estimates, based on ordinary shortwave experience, are misleading because of the terrific competition you expect from other hams, commercial stations and radio frequency noise.

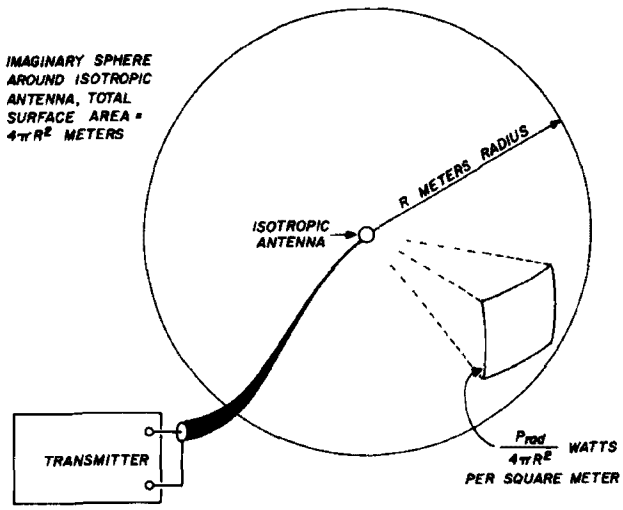
Finally, although the over-all problem is complicated indeed, **the elements can be attacked one at a time.** We can do this in a style resembling elementary accounting: if we can grasp one key point, the rest of the solution will follow.

power

What is it that goes from power supply to transmitter, to transmitting antenna, through space, to receiving antenna, and at last to the receiver? It's power. Voltage and current merely indicate the presence of the power, and calculations with these quantities

represent avoidable detail. If we confine our attention to the really basic central matter of power, we can trace the power through its entire route and analyze each factor causing a change in power level. The power at the end of the journey is fed into the receiver. Let's imagine we are examining a transmitting system.

fig. 1. Classic principle for measuring watts per square meter.



transmitting through an isotropic antenna

We open our attack on the question of received power by imagining we have a transmitter connected to an isotropic antenna. The antenna efficiency is 100 percent. By definition, the power is radiated equally in all directions. There is no such thing as a real isotropic antenna you can hold in your hand, just as you can never manufacture an equivalent circuit to observe the real circuit it represents. Isotropic simply means, "equal properties in all directions," a very convenient imaginary property. All directions would receive equal amounts of rf, or lack of it, if you are a vhf enthusiast inclined to view things from a perspective of large-array behavior.

Now let's go to the anodes of the transmitter output tubes. This is where our dc input power is converted into rf power. The quantity of rf generated may be a little uncertain, but simple voltage and current measurements will give an excellent power input figure. We can use this, as a first approximation, by assuming some reasonable efficiency figure, say 60 percent.

The newly manufactured rf power starts off to the antenna, but not all of it arrives. Some is lost in the transmitter output circuit, and some heats up the transmission line or antenna. The remainder is radiated into space.

Handbook research, careful measurements, ballpark measurements, or just plain thinking can yield values for the losses, which we can describe as unwanted resistances. We're interested in these estimated values because we want to reduce losses, and because even though we may be only estimating sometimes, a carefully thought-out estimate is far better than one that comes off the top of the head.

Now, let's suppose our transmitter input power is 100 watts, an rf-ammeter indicates about 1.1 amperes into a 50-ohm line, and our swr meter indicates zero reflected power from the isotropic antenna. What can we do with this?

Current into a nonreactive, properly terminated transmission line relates to power without a power-factor correction. Therefore I^2R equals about 61 watts going out to the antenna. Call it 60 watts. We can assume a loss resistor in the transmitter, whose influence is appropriate to 26 ohms. Part of this resistance is derived from the power tubes, and the rest resides in the matching network. We may wonder about antenna loss resistance, but typically this is quite small, and for now we're assuming an isotropic antenna anyway. So at this point we have about 60 watts going out into space, and our thinking about its origin would work for any other transmitter, whether tube, transistor or varactor output stage.

the radio-frequency space wave

A large proportion of the engineering difficulties found in real-life work and briefly listed earlier are related to sky- and ground-wave propagation behavior, which becomes decreasingly important as we increase transmitter frequency into the ranges used for space communications. If we suppose there is no propagation other than by space wave,

these complicating factors disappear, yet we still have a usable way of looking at things. We'll choose the simplest workable approach, and when we understand it, we can always fit in some more factors if we feel the need. For now I'll ignore them completely.

Suppose our transmitting antenna is placed inside a huge balloon, or a perfectly nonreflecting sphere many wavelengths in diameter (see **fig. 1**). No interaction will occur between the sphere and the isotropic antenna at its center. Knowing the radius of the sphere, how do we estimate how much rf is striking any part of it?

We choose a system of dimensions first. I prefer to work in meters, because this system is more convenient than the English system, and because we already reckon wavelengths in meters. This will be convenient later, when we come to receiving antennas. You may find it handy to remember a yard is 10 percent shorter than a meter, or that 10 meters very nearly equals 11 yards.

Our sphere has a radius of R meters. A quick reference to a handbook tells us a ball of R meters radius has a surface area of $4\pi R^2$ square meters. Knowing that each square meter of the ball's inside surface receives an equal amount of rf power, we write a quotient for watts per square meter. It comes out as $P_{rad}/4\pi R^2$ watts per square meter, where P_{rad} is the total radiated power in watts.

The sphere is required only as a convenient mental image to remind us how we obtained this watts-per-square-meter result. We can say R represents range rather than radius. Now we can reckon the power striking a square meter of area at any distance, using the same equation. For example, suppose we move out to 10 miles from our isotropic antenna, still radiating an estimated 60 watts. We have a sheet of paper 1-meter square, or about 39.4 inches on a side, and we face it directly towards the transmitting antenna. How much rf is striking it?

A mile is about 1600 meters, so the range is 16,000 meters (see **fig. 2**). The result is 1.9×10^{-8} watt per square meter striking one square meter. At one-half the distance the incidence power would be four times greater; at three times the distance nine times less, etc.

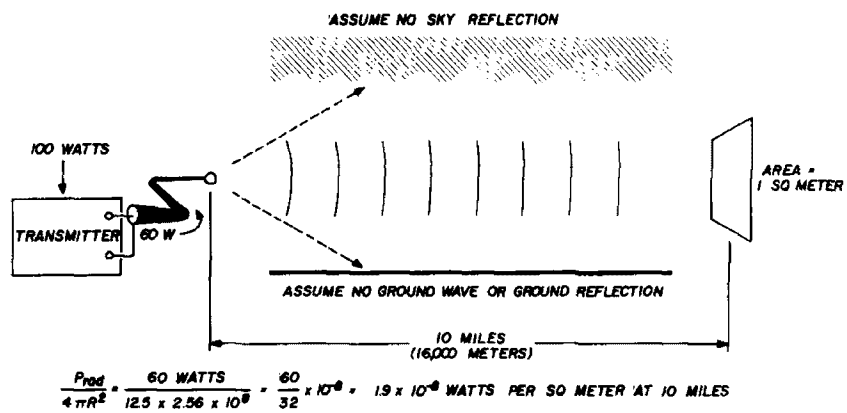


fig. 2. Estimating the space-wave power level at a specific distance. (Note several simplifying assumptions.)

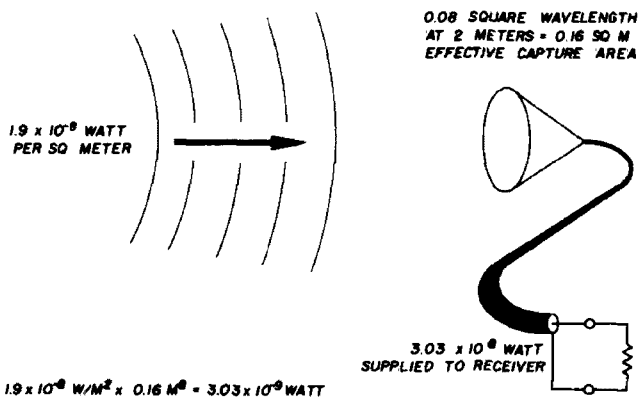
receiving through an isotropic antenna

Now consider the question of how much power can be collected and fed to the receiver. To avoid complexities arising from antenna properties, imagine another isotropic antenna connected to the receiver. How much power will the isotropic antenna intercept?

Since we know the field strength in watts per square meter, we only need know how many square meters of capture area the antenna effectively presents to the incoming wave. A useful list[1] shows the effective capture area of an isotropic antenna to be 0.08 square wavelength. Converting this to meters, and multiplying by watts per square meter, provides our estimate of power available to the receiver antenna terminals (see **fig. 3**).

For instance, suppose our 100-watt transmitter, situated 10 miles away, is operating at a wavelength of 2 meters. Our isotropic antenna at 0.08 square wavelength has an effective capture area of 0.16 square meter, so that it delivers 1.9 x 10^{-8} watt per square meter, times 0.16 square meter, or about 3.03 x 10^{-9} watt to the receiver. This completes our computation by the detailed method of received power if we know transmitter power and range and use the simplest possible antennas. The approach I've used so far is a very good one to understand, but in real life it's rather slow and clumsy and doesn't yield results in the most meaningful form.

fig. 3. Estimating received power.



computing power using decibels

The decibel system often seems confusing, because too many writers neglect to mention their standard unit of power or voltage. Another difficulty arises when we forget that dB numbers represent ratios. When we add dB's we use a shorthand method for multiplying the ratios. For instance, zero dB may look a little odd until you remind yourself it means output equals input, rather than output equals zero. Decibel values of various ratios are contained in many engineering books. Ballantine's handy dB/Ratio conversion slide rule is even better, and I used mine for the dB calculations appearing a few paragraphs below.

Every power transformation or loss factor in our transmitter-receiver system from power supply to power-tube anode, to transmitting antenna, to space, to receiving antenna, and finally to the receiver terminals, can be expressed as a ratio. The transmitter efficiency is P_{out}/P_{in}, the watts per square meter is expressed by $P_{rad}/4\pi R^2$, and so on. Each of these can be expressed in its dB value. The result of all these multiplications and divisions is obtained by adding algebraically all positive and negative values. Since the received signal is much weaker than the transmitted signal, the negative values have to win by a sizable margin. Let's work out our transmitter-receiver problem using dB's.

Here it is again for convenience:

1. 100 watts dc input.
2. 60 percent efficiency.
3. Wavelength 2 meters.
4. Isotropic transmitting antenna.

Ten miles away another isotropic antenna is connected to a receiver. Orientation of isotropic antennas is not important. How much *rf* input power is fed to the receiver antenna terminals?

Since we're talking about power, we'll choose a standard power level. This is 1 milliwatt. We're not concerned with circuit impedances or with ac, dc or frequency. I'll avoid the usual reference standard difficulty by using the term dBm to refer to all power levels. Plus values represent powers greater than 1 milliwatt; minus values represent power levels between zero and 1 milliwatt. We start with dc power at the tube anodes.

Meter voltage and current readings indicate 100 watts input, which is plus 50 dBm. Some is lost in the transmitter, with theory and experience suggesting an efficiency of 60 percent. 100/60 equals 1.67. Referring to our dB rule, we find this is a loss of about 2.2 dB. At this point our power level is reduced to plus 47.8 dBm. This is the power radiated into space.

Since I want to get the same result by dB calculation as by the direct method, I can't throw in any new loss factors. But if we wanted to include an efficiency of 86 percent for the antenna, we would take off another 1.1 dB. I'll pass that by this time, however.

At this point we have plus 47.8 dBm going into space. Our filling-all-space geometry brings in the $1/4\pi R^2$ factor, which consists of a constant $1/4\pi$, and a $1/R^2$ range factor. The $1/4\pi$ is a fixed minus 11 dB, and the range factor requires a manipulation deserving close attention. To calculate the dB value of $1/R^2$ for 16,000 meters, we separate the range number into two parts, convert the parts to dB values, add these, then double the result to represent the squaring. That is, $1/16{,}000^2$ equals $(1/16 \times 1/1{,}000)^2$. Ballantine's handy rule gives us minus 12 dB for the 1/16, and minus 30 dB for the 1/1,000. The sum is minus 42 dB. Doubling this brings it to minus 84 dB. (This is simply calculating with logarithms.) Now our power calculation is plus 47.8 dBm radiated, minus 11 dB geometry factor, minus 84 dB range factor. This yields minus 47.2 dBm per square meter at the receiving antenna.

Our isotropic receiving antenna has a capture area of 0.08 square wavelength times a wavelength of 2 meters, or 0.16 square meter. Our Ballantine rule converts this to minus 8 dB. Minus 47.2 dBm per square meter, minus 8 dB for effective capture area, adds up to minus 55.2 dBm fed to the receiver. If the capture area had been 1 square meter, we'd have used 0 dB here for minus 47.2 dBm at the receiver, etc.

table 1. The "accounting" method for finding received power.

parameter	effective power gain	effective power gain
transmitter dc input power	50 dBm	
transmitter efficiency (60 percent)		—2.2 dB
transmission line efficiency	0 dB	
antenna efficiency	0 dB	
antenna gain over isotropic	0 dB	
spherical geometry factor		—11 dB
range factor		—84 dB
propagation factor		0 dB
isotropic antenna capture area gain (frequency dependent)		—8 dB
antenna gain over isotropic	0 dB	
antenna efficiency	0 dB	
transmission line efficiency	0 dB	
receiver input efficiency	0 dB	
	50 dBm	—105.2 dB

check calculation:
50 dBm — 105.2 dB = —55.2 dBm (accounting method)
3.03×10^{-9} watts = —60 dBm + 4.8 dBm = —55.2 (geometry method)

table 2. Estimates of receiver input power from several antennas.

antenna system	receiver input power (dBm)	equivalent voltage across 50Ω load (μV)
isotropic	—55.8	360
2 dipoles at 2.2dB/dipole	—51.4	600
dipole transmitting antenna/2.2 dBi	—40.4	2100
plus Yagi receiving antenna/13.2 dBi		
Yagi transmitting antenna/13.2 dBi	—31.4	6000
plus colinear receiving antenna/11.2 dBi		

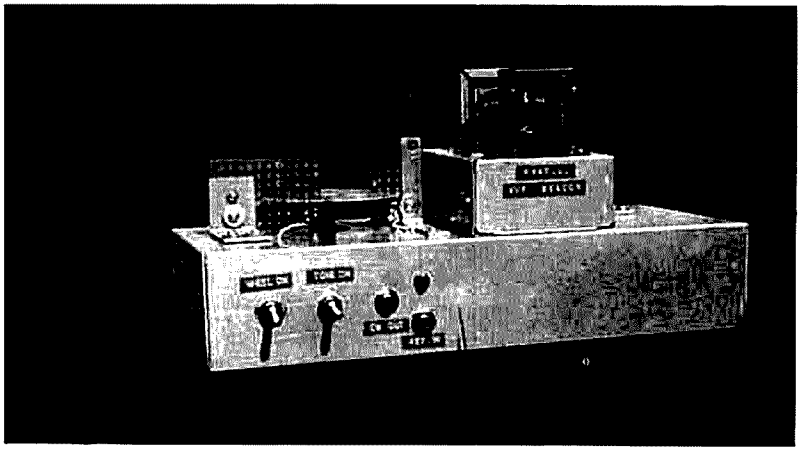
Note: dBi stands for dB gain over isotropic antenna

It's laborious and confusing to try to work out all calculations in the broken-up style I used to try to present these ideas. If we take a hint from the bookkeepers we see this is exactly an accounting problem. We can rearrange it in the neat, concise style of **table 1.** This arrangement offers instant comprehension. It is especially valuable in experimental work where you may be looking for the best way to increase received power. See **table 2** for a sample analysis of results achieved by changing antennas. Such an investigation on paper can be carried out with the assistance of a few simple tests to improve your perspective on actual system performance in your location. You should expect to save many tens of dollars and hours of hard work. Perhaps you can find business applications for it too, with tv applications at the head of the list.

reference

1. "Reference Data for Radio Engineers," IT&T, 4th ed., p. 750.

ham radio



an automatic two-way dx beacon for vhf

Simple method to ensure that you'll be there during band openings

Bob Cooper, Jr. KV4FU/K6EDX

Operation on the vhf bands for the serious DX enthusiast is lonely at best when he's separated by some distance from others with similar interests. Located as I am in the Virgin Islands, some 1100 miles from the mainland, I used to spend a lot of time listening to antenna noise—until I built the DX beacon described in this article.

With 50-MHz conditions building up to a sun-spot cycle 20 peak, I'd been giving some serious thought to how I could be sure that, when the band was open, I would be **listening.** Often I could get on the air if I knew when to be there. Many times I'd be listening, and the other fellow would be listening, but we'd listen right through the opening because we heard nothing. Clearly, I needed something that would make noise (transmit) and alert me when my transmission attracted someone else's attention.

I'd just completed the DX beacon described here when I got into a round table discussion with two other "vhf hermits," OA4C and CE3QG, both of whom had also given some thought to the problem. All of us wanted a beacon to attract attention, but none of us had a ready supply of parts. In each case, the resulting beacon was built from what was on hand. OA4C used what he calls a "promotion motor" to energize his transmitter. (It was promoted from a local

liquor store display for beer.) CE3QG used a tape loop. These approaches are discussed later.

My system is a little more complex than either of the other two. I built it around a spare 6C4 tube I had on hand and the other not too pretty but serviceable components shown. The KV4FU system is described in some detail; those of OA4C and CE3QG in a general way. Whichever approach you take, chances are you will modify it to suit available parts.

KV4FU system

The secret of making the package **two-way** is to use the vox circuit in your ssb transceiver or transmitter. My basic system is shown in **fig. 1.** The code message is etched onto a wheel made from a phenolic board. A 1-rpm motor rotates the wheel past two fixed contact fingers. The center finger makes contact with a copper ring etched around the center of the wheel. The second finger contacts the etched code characters around the periphery of the wheel. The make-or-break circuit carries an 800-Hz sine wave output from an audio generator, which is fed through a bandpass filter. This signal drives the transmitter (or transceiver) through the vox circuit. My message, shown in **fig. 1,** lasts 35 seconds. The remaining portion of the wheel is blank, which is a 25-second listening period. After the "K" is transmitted, the vox circuit returns the transmitter to the receiver mode. This is when other stations can call or make noise to alert me of their presence.

code wheel construction

I made my code wheel from a 4 x 6-inch single-sided phenolic board (phenolic on one side only).* The code characters can be made with EZ Etch pressure-sensitive transfers, or you can use 1-point and 2-point ruleline tape, which has an adhesive backing. This is available in art supply stores.

Lay out your message in pencil. After scribing the circle for the wheel with a compass, center the copper ring for the fixed contact. Etch the code characters, then cut out the wheel circle with a jig saw or coping saw. Drill a hole exactly in the center for a press fit onto the motor shaft.

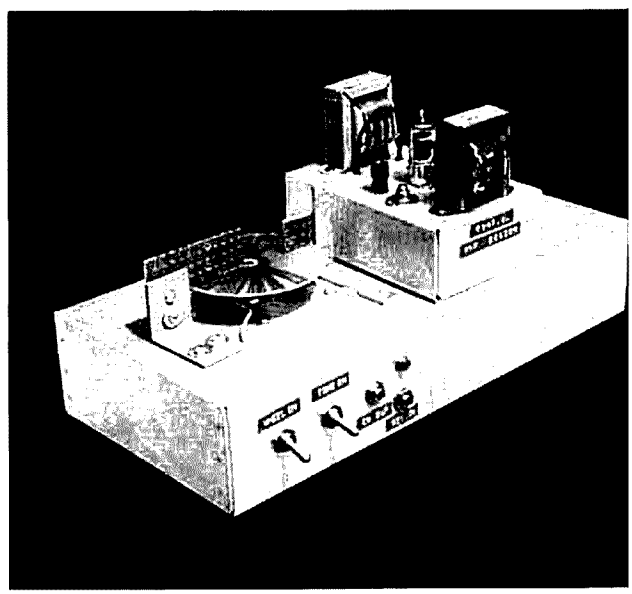
* Available from Amidon Associates, 12033 Otsego Street, North Hollywood, California 91607. They also handle EZ Etch products.

keying mode

Since you are keying an ssb transmitter with an audio note, the transmitter input will be single-tone audio and not pure cw (depending upon the method used to generate cw in your particular unit). The **purity** of the audio note originating in your 800-Hz audio generator will, to a large measure, determine what your beacon sounds like at the other end. Take care not to overdrive the audio input of your transmitter with the relatively high output levels available from most audio generators (i.e., don't flat top your cw).

I have experimented with leaving the 800-Hz audio bandpass filter out of the system with varying results. When the distortion products from the audio generator are held to levels 50-60 dB below the 800-Hz output, there's no need for the audio bandpass filter. In my own receiver, however, the unwanted (lower) sideband is still present and objectionable, although it is at least 45-50 dB below my desired (upper) sideband transmitted signal. A slightly less-pure 800-Hz audio output produces a combination of cw and mcw,

The KV4FU DX beacon. Contact fingers for code wheel are attached to phenolic boards mounted at right angle to wheel. Slit in boards provides track for wheel.



which makes it possible for stations without a bfo to copy your beacon. (Some stations in South America are still so equipped on 50 MHz.)

When the signal is quite strong, audio notes will be present next to the 800-Hz signal and probably won't contribute to your communications range or win you compliments from those hearing your signal. When signals are weak, the cw portion (driven by the stronger 800-Hz note) will probably be all the distant station will hear.

the audio generator

The circuit I use is shown in **fig. 3.** The audio output transformer, L1, is a Stancor A7949 whose 2000-ohm primary is the series element in a pi network. The secondary is not used.

I chose 800 Hz because I had some surplus toroid filters on hand, and this frequency complemented the cw operation of my Swan 250 transmitter. The Swan's cw signal is generated 800 Hz above the receiver listen frequency, and a station hearing my beacon and zero beating with another Swan 250 (let's face it—there are a lot of them around) will also produce an 800-Hz audio note in my receiver. Your choice may be different for similar reasons.

The response of the surplus filter is quite good. A scope showed sidebands of 750 and 850 Hz to be down 30 dB or more. The 1968 edition of the ARRL handbook (pages 50-51) contains data on building your own filter.

power supply duty cycle

Some attention should be given to the duty cycle of your transceiver power supply if you intend to run extended operations with your beacon. Exact data on the Swan 117XC duty cycle is not available to me, but it appears that a 25 percent duty cycle does not overtax the unit. I operate on transmit mode 60 percent of the time, but I transmit only a portion of this period. The end result is that my key-down time is 25 percent of each minute. I have experienced no difficulties with the Swan 117XC.

vox circuit adjustment

The vox circuit delay control should be adjusted so that the transmitter stays on the transmit mode (i.e., remains keyed with resting current present) between characters. This will ensure that you don't clip the leading edge of characters and distort the sound of your cw.

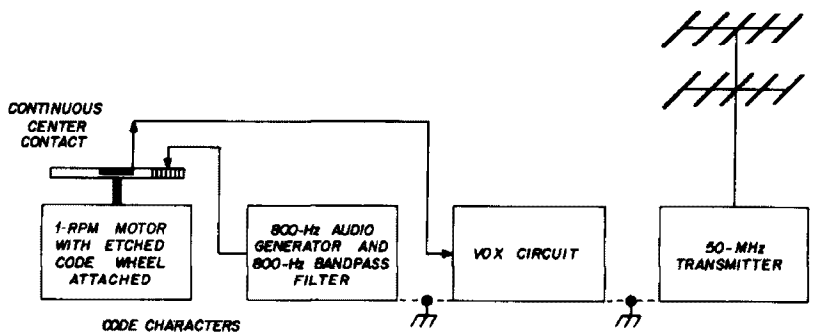


fig. 1. Block diagram of KV4FU vhf DX beacon. Microphone input to transmitter is driven through vox circuit via code wheel, which keys audio generator output.

rf feedback

You might experience some problems with transmitter rf getting back into the audio generator system and keying circuit. I found it necessary to physically remove the audio generator code wheel about six feet away from my 50-MHz final amplifier to prevent rf distortion of the audio keying signal. The code wheel output is coupled to the audio input jack on the Swan 250 through a shielded microphone cable. (A transistorized audio generator had to be enclosed in a shielded box and a copper screen built around the code wheel in a second unit I constructed. For this reason, the tube unit shown here is now in use, because it was less susceptible to rf overload.)

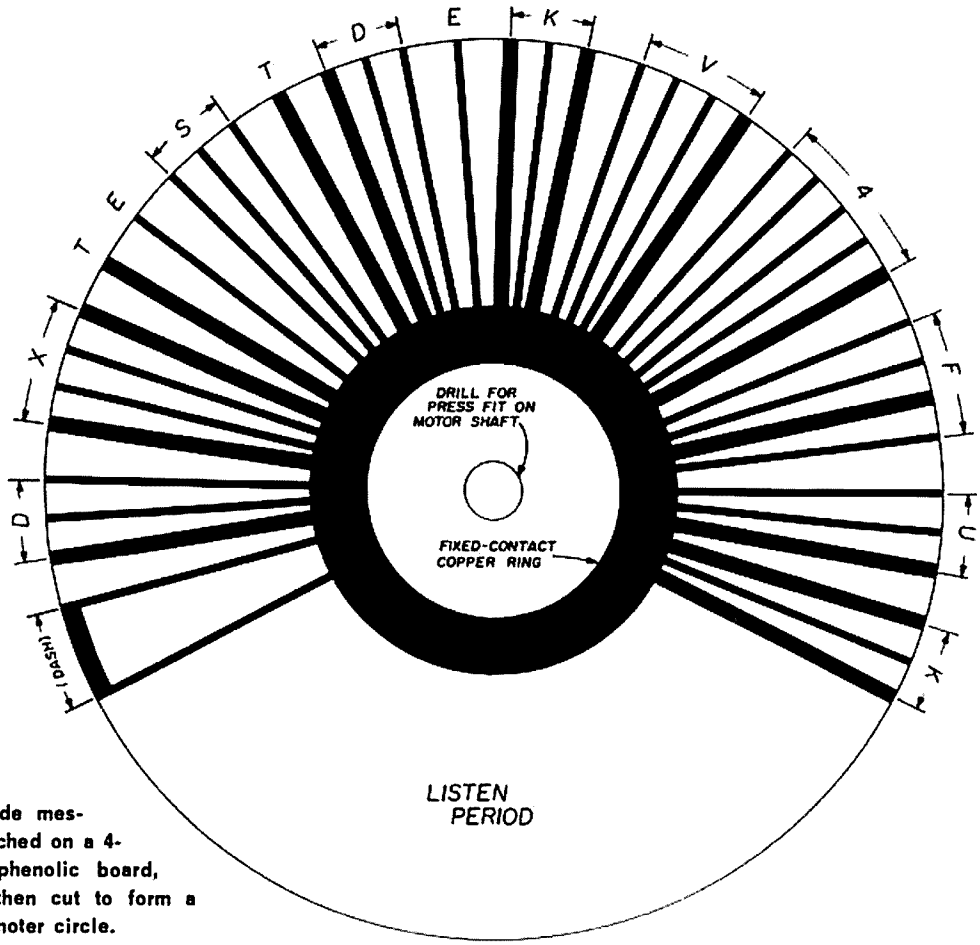


fig. 2. Code message is etched on a 4- x 6-inch phenolic board, which is then cut to form a 4-inch-diamoter circle.

The photos and diagrams should provide adequate explanation for anyone interested in duplicating my DX beacon.

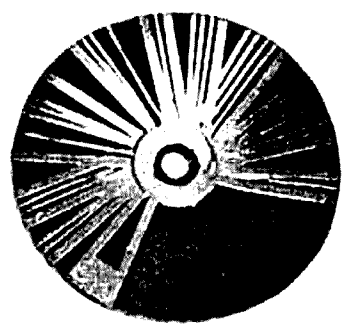
CE3QC and OA4C approaches

The CE3QC method of accomplishing the same objective is similar, and OA4C has already updated CE3QG's method with another step.

CE3QG prerecorded his beacon message on an audio tape loop which consists of "(dash, dash) CQ TEST CE3QG TEST CE3QG." The tape loop is recorded in mcw, and the audio output from the playback head on the recorder drives his Swan 250 VOX circuit. One advantage to this method is that beacon messages can be changed quickly to suit his needs without having to etch a new code-wheel board. The delay period and timing can also be changed at will.

OA4C liked this idea and expanded upon it by including an aural CQ on his tape interspersed with his cw message. He sends (on cw) "CQ CQ DE OA4C," followed by "CQ

The completed code wheel. Characters can be made with pressure-sensitive transfers or tape available at art supply stores.



SIX METERS FROM OCEAN ABLE FOUR CHARLIE LIMA PERU" on ssb.

Heinz, OA4C, also has a band watching mechanism that alerts him when signals appear on the 50-MHz band. He has only one other local, OA4BR, who cooperates by not falsely triggering Heinz's warning system. Here is how it works.

the OA4C band watcher

Heinz' "promotion motor," which is ac operated and self reverses its shaft rotation

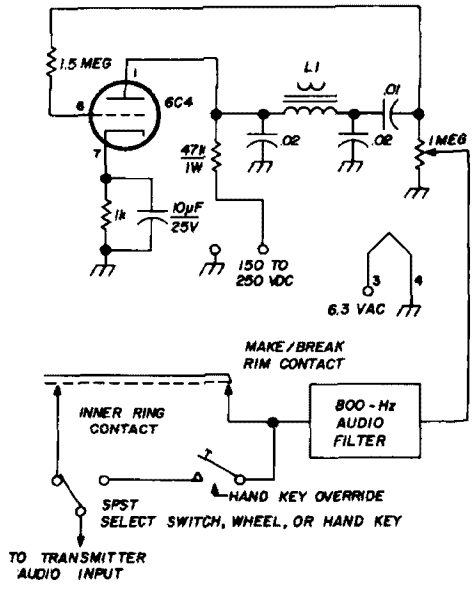


fig. 3. DX beacon audio circuit. Output transformer is Stancor A7949 primary (2 kilohms); secondary is not used. Audio bandpass filter is surplus kit.

after one complete revolution, is connected to his receiver tuning dial through a gear arrangement. The motor remotely tunes his receiver plus or minus 50 kHz from center frequency. He places his microphone in front of the receiver speaker, sets his vox sensitivity circuit so that any received signal will key the vox, and throws a spst switch that disables the transmitter vox relay control. This allows the vox relay to ring a bell. When he hears the bell, he knows a DX signal is on the band, and he hotfoots it back to the shack.

are beacons really necessary?

Many years ago numerous 50-MHz beacons were on the air. Industrious amateurs put them on to warn of DX conditions, and the beacons ran almost continuously. As activity grew through sunspot cycles 18 and 19, the beacons slowly left the air on the theory that with much activity one didn't need a beacon to tell one when the band was open.

This is all well and good for DX openings via the E layer within the Continental United States. However, anyone who has ever listened to commercial signals just below (or within) the 50-MHz band during periods of high muf via the F layer will tell you that the band is open for hours on end but amateur signals are not present. The need for beacon activity both inside and outside of the Continental United States has never been greater. As CE3QG, OA4C and others will tell you, stateside 49-50 MHz signals are often heard, but with no signs of amateur activity.

A similar case exists for two-meter E-layer DX. The construction of a vox-controlled beacon for two-way vhf activity overcomes the old bugaboo about beacons being continuously run and the operator transmitting right through a DX opening when he should have been listening. With such an installation, more contacts will result, even in times of short-period openings and rapidly changing conditions.

Vox-controlled ssb transceivers abound on both six and two meters. In just a few hours you too should be able to put your own beacon on the air. And listen for mine on 50.1028!

ham radio



"If you'd read "ham radio magazine" carefully, you'd know you heat the wire and not the solder, dear."

high-linearity voltage controlled crystal oscillator

Peter Laakmann, WB6IOM, 8001 Airlane Avenue, Los Angeles, California 90045

While working up plans for a special weak-signal receiver with precision calibration, I needed a stable vco (voltage-controlled oscillator) that could be tuned over a small frequency range—10 kHz at an output frequency of 28 MHz. The oscillator had to have good voltage-vs-frequency linearity as well as crystal stability. With good linearity the frequency could be controlled with a ten-turn potentiometer; the ten-turn dial would then read frequency at the rate of 1 kHz per revolution, and setting repeatability would be better than 10 Hz with calibration accuracy of 50 Hz.

The weak-signal receiver is not finished yet,* but it will use direct conversion of the 28-MHz i-f signal to audio using ssb phasing techniques for image rejection. The vco feature is needed for automatic search modes and phase-lock inputs:

The voltage-controlled crystal oscillator (vcxo) worked so well I decided to describe it separately since there are many applications for such a unit. It could be used as the master oscillator for a vhf transmitter (resulting in considerable tuning range of the output frequency) or it could be used to generate wideband fm in a vhf transmitter or fsk in a lower frequency unit. These are all applications that are difficult to accomplish with conventional techniques if extreme carrier stability is required.

The electronic tuning of a vco permits some interesting tuning concepts for receivers and transmitters. For example, more than one dial can be used with a vco, and the dials can be switched electrically. With this arrangement the vco can "remember" a particular tuning position, or it can be tuned off frequency with the second dial. In a transceiver a vco can provide independent tuning for either transmission or reception. In a moonbounce system a vco can be used to remove Doppler shift from the expected echo by changing vco frequency by the expected Doppler shift between reception and transmission.

the vcxo

This oscillator is designed around a crystal that has a series-resonant frequency at 14 MHz. Crystal frequency can be shifted with good linearity over a range of plus or minus 10 kHz at 14 MHz. My application only calls for ±2.5 kHz deviation at 14 MHz; over this range linearity is within 0.25 percent. From my experiments with this circuit it appears possible to shift the frequency as much as 30 kHz.

The vcxo is temperature compensated and has a frequency stability of 2 Hz per degree centigrade. Temperature compensation is required because the frequency is influenced by circuit parameters other than the crystal. Stable operating voltages are also required for the same reason. Stability of the completed vcxo is better than the best LC oscillators by at least an order of magnitude; **fig.**

* Author WB6IOM promises a complete rundown when the receiver is complete.

1 shows frequency vs temperature, operating voltage and control voltage.

Output level of the circuit is 1 Vrms sine-wave into a capacitive load of 10 pF. The output circuit is **not** designed to operate into a 50-ohm load because this would increase power dissipation inside the unit and cause undesirable warm-up effects.

The circuit settles right down when the frequency-control dial is reset—frequency changes are instantaneous. Warm-up drift is about 10 Hz, with stabilization to 1 Hz within 10 minutes. The oscillator will stay within

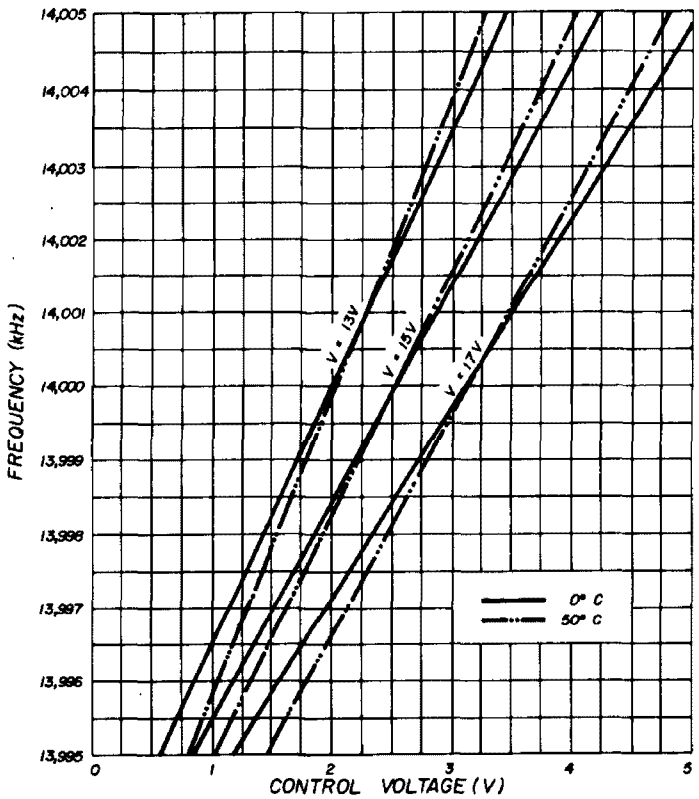


fig. 1. This curve of vcxo frequency characteristics illustrates the excellent linearity and voltage and temperature characteristics of the unit.

1 Hz for hours in a normal room-temperature environment as measured with a digital counter.

operation

The circuit shown in **fig. 2** uses the crystal in a series-resonant mode between a common-base stage (Q1) and a common-collector stage (Q2). While it might seem that the impedances of both stages are quite low, the crystal actually sees the impedance of the emitter resistors rather than that of the transistors. This is because the amplifiers are not operated class A due to the large amount of loop gain that is available. The large loop gain causes the oscillation amplitude to build up to a level where the transistors are cut off for a considerable portion of the cycle.

The series-resonant impedance of the crystal is approximately 10 ohms, so the large series emitter impedance reduces the effective crystal-circuit Q by a factor of 100 or more. The reduction of crystal Q allows amplifier phase shift to change the frequency of oscillation. Most of the initial amplifier phase shift is produced by the large collector resistance of the common-base stage and its collector-to-base capacitance. This causes the frequency of oscillation to be lower than crystal resonance.

The variable capacitance of the 1N4452 diode produces a variable phase shift that can raise or lower the frequency of oscillation. Although this diode is not characterized as a varactor it is used as one in this circuit. I have used many of these diodes in various applications as harmonic multipliers. Typical minus 6-volt cutoff frequency is 20 GHz. Zero-bias capacitance is typically 30 pF; at 20 V it is a few pF. Other diodes with similar parameters will probably work too, although regular switching diodes do not work.

In this circuit the diode is forward biased over part of the rf cycle; the control voltage varies the conduction angle of the diode. This probably explains the exceptional linearity. As soon as the control voltage is raised above the point where the diode current goes to zero, linearity rapidly deteriorates.

At high forward diode currents the oscillator frequency is below the resonance of the crystal because of the large phase lag in the common-base stage. At low diode currents (high control voltage) the average capacitance of the diode increases the effective crystal resonant frequency. Oscillator frequency is then above the crystal resonant frequency.

Since the collector resistance of the com-

mon-base stage controls phase shift, changing its value with temperature can effect the temperature compensation of all other temperature-sensitive elements in the circuit. This is accomplished by the unijunction transistor (Q4) series resistor combination that shunts the collector load resistance. The base diode of the unijunction transistor is not connected. A "Sensistor" will do the same job but it is more expensive and is essentially the same piece of silicon anyway. I suspect that each oscillator needs to be compensated individually; the 12k series resistor can be varied to compensate properly any

tuneup

There should be no problems with crystal activity in this circuit since feedback is quite large. The linear tuning range can be centered around the desired center frequency by varying the common-base collector load. Decreasing the 12k series resistance to the unijunction transistor will increase the temperature sensitivity of the load resistance and can be used to obtain exact temperature compensation for a particular control voltage. Temperature compensation will then be close enough at either end of the tuning range.

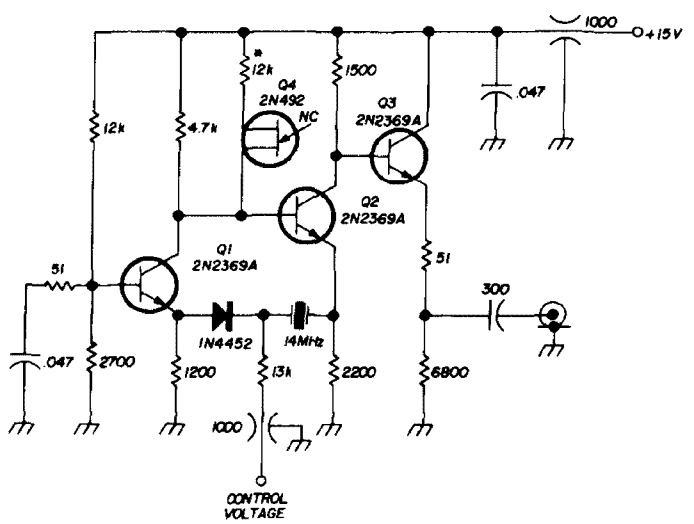


fig. 2. Schematic for the voltage-controlled crystal oscillator. The 12k resistor in series with the unijunction transistor is used for temperature compensation. The crystal is a CR-19/U type available from Texas Crystals.

oscillator. More about this under the section on tuneup.

I used 2N2369A transistors but any other high-speed switch or amplifier type should work equally as well. The output stage is an emitter follower. This stage does not provide complete isolation of frequency from load variations but is sufficient for most uses.

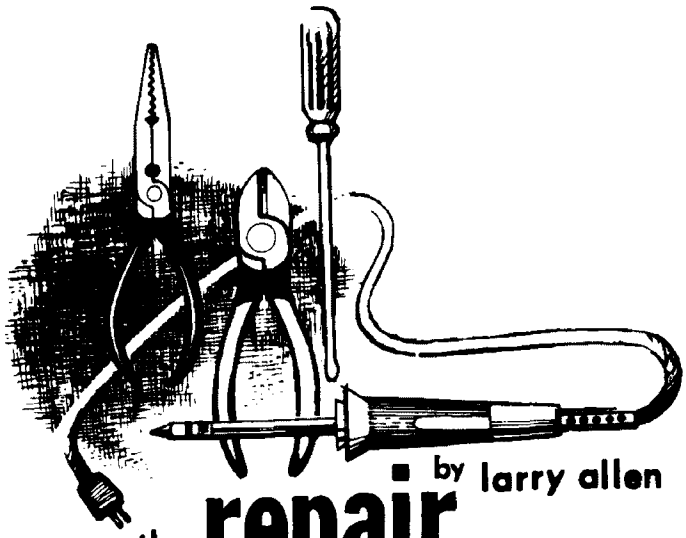
construction

The oscillator is constructed in a small metal box, 1½ x 1½ x 2½ inches, that is completely closed to maintain a uniform temperature. Dc voltages are brought through feedthrough filters. Output is from a BNC fitting. The oscillator can drive a few inches of coaxial cable without much loss in voltage. All resistors used in the prototype are ⅛-watt metal film types. These provide long-term stability.

Decreasing the 13k resistor in series with the control voltage will yield increased tuning range toward lower frequencies. Decreasing the 4.7k collector resistance of the common base stage is the only variable available to control the high end. If the oscillator is used with higher supply voltages, slightly more tuning range may be obtained.

The oscillator should be used with a stable power supply. The temperature compensation of the oscillator assumes that the power supply used has a near zero temperature coefficient of voltage. If the power supply is located in the same thermal environment it can, of course, be included in the thermal compensation of the oscillator. Ripple on the power supply must be less than a few hundred microvolts, particularly on the control voltage line.

ham radio



by larry allen

the repair bench

tuning up ssb transmitters

On a quick look, tuning up a single-sideband transmitter may appear about the same as tuning any other kind. But when you try it, you'll find there are some considerable differences.

In fact, a complete ssb tuneup is something suited more to the bench than to the working shack. A full ssb tuneup is really an alignment job. There are critical adjustments to be made. Fortunately, once made, these screwdriver or transformer-slug adjustments can be left alone as long as the set works okay. You don't have to change them every time you change bands or frequencies.

The chief trouble most hams run into in aligning an ssb transmitter stems from not really knowing what each step does. If everything happens according to the book, okay. The trouble is, not everything does. It's then that knowing and understanding determines how the job turns out. This background of know-how and know-why is what I plan to bring you this month. Then you'll be at ease with the nuts and bolts of ssb transmitter alignment I'll tell you about next month.

ssb band-changing

Remember that the main carrier oscillator in an ssb transmitter does not determine output frequency. You can see why in **fig. 1.**

The carrier oscillator generates its signal at a fixed frequency—sometimes 455 kHz, and in transceivers, usually around 5 or 9 MHz. This signal has only one purpose: creating sidebands. The voice signals need some kind of carrier to beat with in the balanced modulator. Otherwise, no sidebands or difference frequencies would be generated.

The output of the balanced modulator is a double-sideband supressed-carrier (dsbsc) signal. The two sidebands contain all the voice modulation, in rf form of course. One sideband is just above and the other just below the carrier frequency.

Then the dsbsc signal is sent through a special filter that allows only one sideband to pass. The output of that filter is a single-sideband signal. It contains all the rf modulation components that were formed on one side of the carrier.

However, in today's equipment, the so-called carrier that is mixed with voice signals to form sidebands is never at the same frequency as the output of the ssb transmitter. With only one signal frequency, the balanced modulator and side-

band filter can be designed for optimum results at that frequency. It would be difficult and expensive to change so many tuned circuits every time you shift transmitter frequency. Tuning up each time would be complicated and time-consuming.

Instead, frequency in a modern ssb transmitter is determined **after** the single-sideband signal is formed. Usually, it's done by a heterodyne method. This is also explained in **fig. 1.**

The ssb signal from the sideband filter is mixed with a cw signal from a master vfo. If the transmitter operates at vhf, there is also likely to be some frequency multiplication between the vfo and the frequency converter stage where the signals are mixed.

The ssb signal and the vfo signal heterodyne and create sum and difference signals. If you figure them out, you'll know that makes a sideband above the vfo frequency by the amount of the original carrier frequency, and another below by the same amount. If the frequency converter is a simple mixer, the output also contains the vfo signal; if it's a balanced mixer, the vfo signal is canceled in the output.

The two new sidebands are quite far apart. If the carrier oscillator is 455 kHz, for example, the two sidebands are 910 kHz apart. Any fairly selective tuned tank can pick out one sideband and eliminate the other quite easily.

The remaining single sideband goes to the antenna. If more power is needed, the signal is amplified first in a linear power amplifier.

alignment fundamentals

Obviously, changing the output band involves only the tuned circuits that **follow** the frequency converter—even if the transmitter has two converters. You switch the tuned tanks. If the band change is from an hf to a vhf band, you might also switch in a multiplier stage after the vfo.

Changing frequency within a band is only a matter of retuning the vfo. You may have to make minor tuning adjustments in the tank circuits that follow the frequency converter. If the transmitter setup includes a linear amp, chances are you tune it up only when changing bands; it's broadbanded.

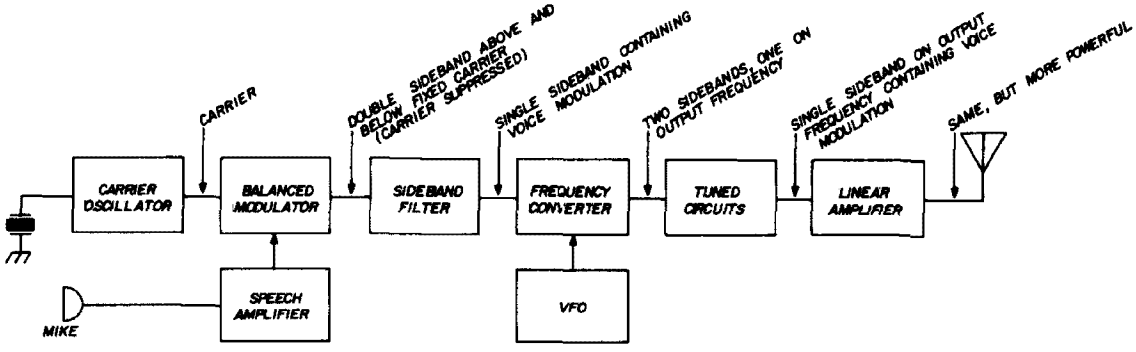


fig. 1. Output frequency of single-sideband transmitter depends on heterodyning already-formed sidebands with signal from vfo, which develops a sideband at transmitting frequency.

However, there are important adjustments to be made periodically if you want to maintain top performance, adjustments that come under the heading of alignment. The balanced modulator is the most critical of these. It must be balanced perfectly to cancel the carrier signal effectively. This entails only two or three adjustments, but they must be accurate and correct. The steps vary from model to model, but their goal is inevitably the same: to suppress the carrier as much as possible.

Sometimes minor—but touchy—adjustments must be made in the vfo. Aging

components and the wear of regular use eventually bring on tuning errors. Alignment restores dial accuracy.

Even stages following the frequency converter have some servicing-type adjustments that need periodic attention. Bias for the linear amplifier is important to operation and to tube life. Some rf stages need neutralizing. Each band usually has separate adjustments that "center" the tuned tanks of the rf output circuits.

peculiarities in transceivers

Certain transceivers use one or more receiver i-f stages to amplify the sideband signal before it reaches the last frequency converter. That ties receiver alignment in with transmitter-adjustment peaking. So, you must align the receiver first. An example of this kind of unit is the partial block diagram of a Hallicrafters transceiver in **fig. 2.** Only the connections to the receiver stages are shown.

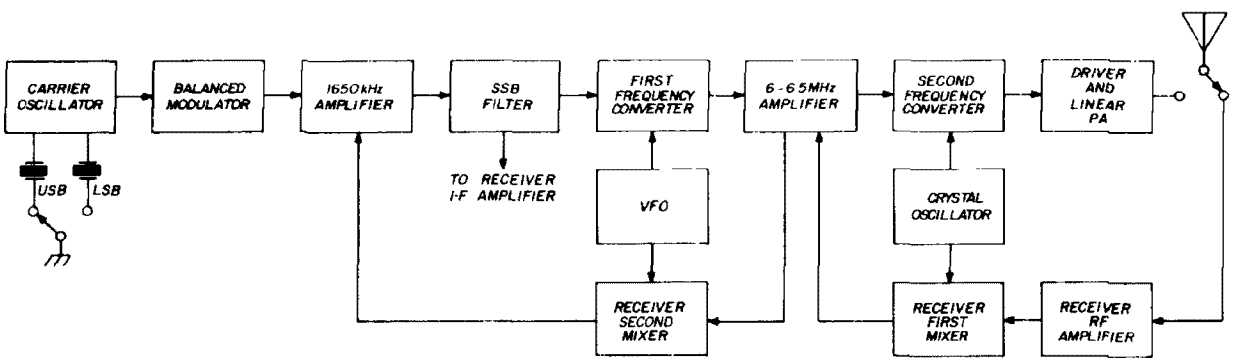


fig. 2. Transceiver uses some of receiver i-f stages to amplify transmitter signals at various stages of frequency conversion. Sideband generated in balanced modulator is heterodyned up to the output frequency.

In the receive mode, the first superheterodyne conversion changes incoming single-sideband signals to intermediate frequencies between 6.0 and 6.5 MHz. The oscillator that feeds that first receiver mixer is fixed-frequency (the same crystal oscillator that feeds the transmitter **second** converter). The signals are amplified by a broadband (6.0–6.5 MHz) i-f amp.

Then they're mixed with a signal from the variable frequency oscillator. Its tuning picks out the one signal you want and converts it to 1650 kHz. From that second mixer, the signal is amplified and fed through a narrowband filter to eliminate any nearby frequencies that might have slipped past the 1650-kHz tuned circuits. Then it's amplified again, detected, and so on. That's the receiver.

Two i-f stages of the receiver are used in the transmit mode, too. Study **fig. 2.** The carrier oscillator uses two crystals, one to generate upper sideband (usb) and one to generate lower (lsb). The usb crystal is at 1652.8 kHz; the lsb crystal is at 1650.0 kHz. The output of the balanced modulator is a pair of sidebands above and below whichever frequency has been selected by the usb/lsb switch.

Bandpass of the 1650-kHz amp is broad enough to pass both sidebands at either frequency. The filter that follows, a crystal-lattice type, has a very sharply defined response. It passes frequencies from 1650–1653 kHz, sharply rejecting any above or below. So, if the usb crystal is chosen, the sideband filter picks out the **difference** or lower sideband from the two generated in the balanced modulator (it's later converted to an upper sideband of the output frequency). If the lsb crystal is chosen, the filter selects only the **sum** or upper sideband of the modulated carrier. The carrier itself is suppressed in the balanced modulator.

The one-sideband signal from the filter is fed directly to a transmitter frequency converter. There it mixes with a tunable signal from the vfo. Again, the sidebands are produced at sum and difference frequencies—the 1.650-MHz sideband mixing

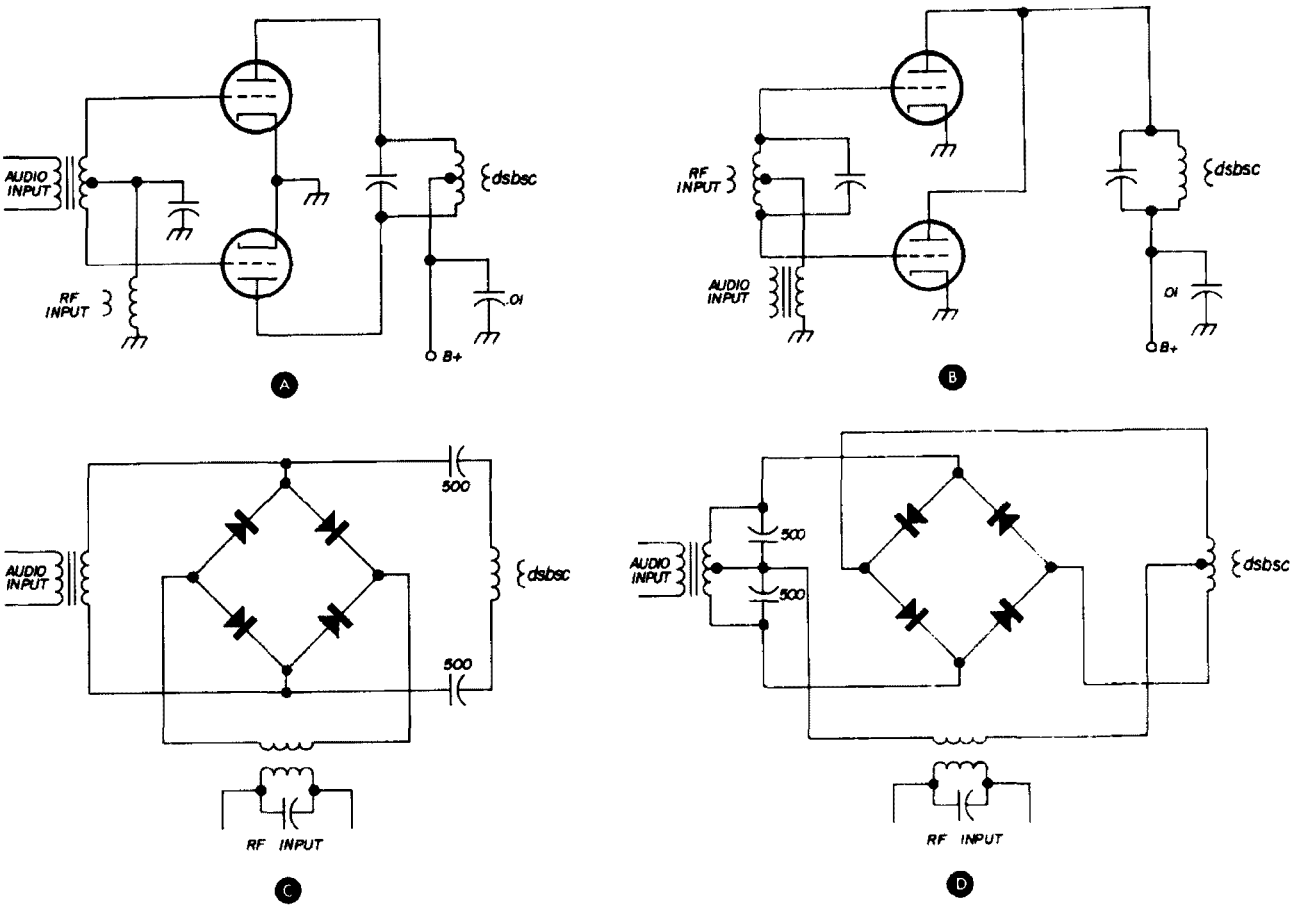


fig. 3. Different balanced modulators you may find in ssb equipment.

with the vfo frequency. One sideband falls in the 6.0–6.5 MHz range; the other falls somewhere around 3 MHz. The 6.5-MHz tuned circuits, in the amplifier that follows, reject the 3-MHz sideband easily.

The desired sideband, between 6 and 6.5 MHz, is amplified and sent to a second frequency converter, a fixed-frequency oscillator. This is where the main band-changing is done. The oscillator has a selection of crystals. With the vfo tuning the signal over a frequency range of 6 to 6.5 MHz, this final frequency conversion determines what band this 0.5-MHz spread is applied in.

For example, a 10-MHz crystal beats with the 6.0–6.5 MHz signals to produce output frequencies between 3.5 and 4.0 MHz, the 80-meter band. A 13.5-MHz crystal gives a 7–7.5 MHz spread—the 40-meter band. And so on. The 10-meter band from 28 to 29 MHz is covered with two crystals, at 34.5 and 35.0 MHz.

inside balanced modulators

These explanations are to acquaint you with exactly what you're doing when you adjust one section or another of an ssb set. One of the least understood stages is the balanced demodulator. I'd better tell you what the adjustments in it are for.*

Briefly, here is the principle of a balanced modulator—sometimes called a balanced mixer. If you feed a signal in parallel to any two perfectly matched stages and take the output from them in push-pull, the input signal is canceled. Or, if you feed the balanced stage in push-pull and take the output in parallel, cancellation occurs.

Suppose the stages are two balanced but nonlinear mixers. You feed them an rf signal in push-pull and voice signals in

*Editor's note: If you don't know balanced modulators at all, you might go back and review them in the article "Generating SSB Signals" on page 24 of the May, 1968 *ham radio*.

parallel. If you take the output of the mixer in parallel, you get only sums and differences; the original rf signal is eliminated.

Examine the stages in **fig. 3.** All four are basic examples of balanced modulators. In **fig. 3A** the rf input is fed to the triode tubes in parallel. The audio is applied in push-pull, and the output is taken in push-pull. The balanced modulator in **fig. 3B** has the audio applied in parallel and the rf in push-pull. The output is taken in parallel. In both configurations, the rf signal is canceled from the output, but modulation (the sidebands) remains.

Diodes are much more popular than tubes in modern commercial ssb transmitters. Semiconductor diodes are smaller, cheaper, and generally more dependable. They don't unbalance as easily as tubes do. **Figs. 3C** and **3D** show the two chief arrangements.

The first, **fig. 3C,** is a bridge. The anodes of two diodes are tied together, as are the cathodes of the other two. The push-pull/parallel rule holds here, too. The audio is fed to the bridge at points that unbalance the diodes ability to conduct. The unbalancing is in step with the audio frequencies. The rf, on the other hand, is fed to the bridge at balance points—the diode branches are alike between them. Rf is eliminated from the output except to whatever degree and at whatever rate the audio input is unbalancing the bridge. The result is a double-sideband output as you'd expect from any diode mixer, but with the rf carrier suppressed.

The **fig. 3D** circuit is called a **ring** modulator. It gets that name from the way the diodes are wired—in series with each other.

In a ring modulator, the output can be taken in either mode, just as long as the rf and audio are applied in opposite modes. In this one, the audio is in the same mode as the output—push-pull. The rf is applied in parallel, being fed in at the center taps of the audio and output transformers. The output is a dsbsc signal.

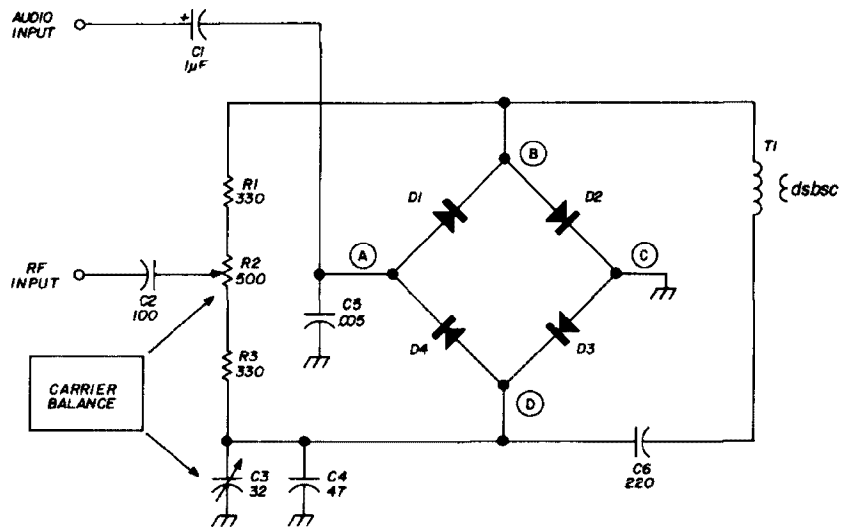


fig. 4. Most popular diode balanced modulator is form of ring type. This one is designed so it can be driven by audio and rf signals from single-ended amplifier stages.

adjusting a balanced modulator

The balanced modulator adjustments are probably the most critical ones in your transmitter. Slight misadjustment deteriorates your on-the-air sound and cuts your PEP output. Yet the principle of what must be done is simple.

The most popular ring modulator today is shown in **fig. 4.** Several commercial sets use it. It's a diode ring modulator, but its arrangement is ingenious.

Points A and C of the diode ring are the feed points for the audio signal. With point C grounded, a single-ended audio signal can be applied to point A.

The carrier signal comes from a single-ended stage, too, but the way it's handled applies it to the diode ring 90° out of phase with audio. (That's in effect what the push-pull/parallel feed does, too.) The rf is applied this way by keeping the output transformer primary and input divider R1-R2-R3 above ground for audio but at ground for rf. Capacitor C4, with trimmer C3, does the grounding. Points A and C in the diode ring are both at rf ground (notice C5).

The output is taken through T1. C6 blocks audio from reaching T1. The double-sideband output signal is coupled to an ssb filter. The carrier that formed the sidebands is canceled.

The adjustments for balance are labeled in fig. 4. The diodes must first of all be pretty well matched; if you replace one, either carefully match the new one to the other three or pick four matched new ones. (You can match them yourself with an ohmmeter. Just be sure that all forward readings are about the same, and that all backward readings are beyond 100k. The forward reading is the most important one for matching.)

Align R2 and C3 with no modulation. Trim them for minimum rf output. The driver is a handy monitor point, usually. Adjust the potentiometer first, then the capacitor. They interact some, so go back and forth a few times.

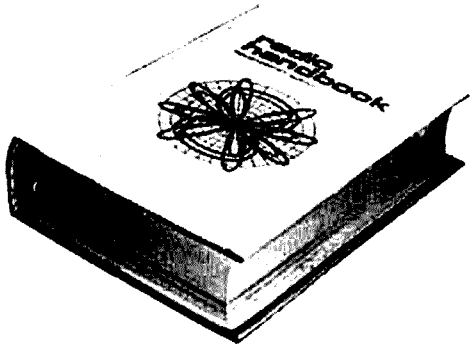
complete ssb alignment

Now you have the fundamentals you need for a proper job of aligning your ssb equipment. If you don't understand what you're doing, you might wind up high and dry if you run into trouble. The job isn't difficult, but it can confuse the uninitiated.

Next month, I'll show you the entire alignment procedures for a couple of ssb transmitters. They are typical ones, chosen because they demonstrate several different kinds of adjustments. I'll also take you through the adjustment of a high-power linear amp, just in case you have one in your shack or one turns up on your repair bench.

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tropospheric-duct communication

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This article is being written with the anticipation that I will have worked KH6EEM on 144 MHz by the time it is printed and that the subject will then be of current interest. I have been greatly interested in tropospheric duct communication ever since the summer of 1957 when John Chambers, W6NLZ, at Palos Verdes Estates, California worked Ralph Thomas, KH6UK, at Kahuku Point, Oahu on 144 MHz. My professional experience on the subject has been with the Radio Meteorology Section of the National Bureau of Standards under Bradford Bean. My Master's thesis from Utah State University, "The Effect of the North Pacific Trade Wind Inversion on Long Range VHF Radio Communications," furnished most of the data for this article. I will assume that you have, or will, read two *QST* articles on the subject [1, 2] and will go on from there.

The troposphere is the domain of the weather. It extends to a height of about 4 to 12 miles and is composed primarily of nitrogen, oxygen, argon, water vapor and carbon dioxide with traces of neon, helium, krypton, nitrous oxide and xenon in descending order of occurrence. With the exception of water vapor the tropospheric constituents are fairly well mixed by convection from the sun-warmed surface of the earth.

The mixture of gases in the troposphere near the surface of the earth is very slightly magnetic with a relative permeability, μ_r, of 1.0000004. The relative dielectric constant, ϵ_r, is much more important for the propagation of radio waves, being about 1.0006. The dielectric constant is greater than unity because an electric field creates a dipole moment in the component molecules by shifting their charge centers slightly from their equilibrium positions. The refractive index, n, the square root of the product of the permeability and the dielectric constant, is about 1.0003 near the surface of the earth.

Since the refractive index is so close to unity it is more convenient to use the refractivity, N, which is the difference between the refractive index and unity multiplied by one million. Refractivity due to the gases other than water vapor is proportional to the molecular density. Water forms a polar molecule, however, which is one with a permanent dipole moment. Ordinarily the water molecules have random orientations and no net field. Under the influence of a radio frequency field the water vapor molecules become partially aligned, vibrating with the rf field, and the net result is an increase in refractivity proportional to the density of water vapor molecules divided by the absolute temperature. The refractivity may be determined from the following formula:

$$N = (n - 1) \times 10^6 = \frac{77.6P}{T} + \frac{3.73 \times 10^5 e}{T^2}$$

where P is the total atmospheric pressure in millibars, e is the partial vapor pressure in millibars, and T is the absolute temperature in degrees Kelvin. The vapor pressure may be determined from the dew point by reference to standard meteorological tables or from the mixing ratio, w, by the following formula,

$$e = \frac{P}{\frac{623}{w} + 1}$$

The form of the refractivity equation is theoretical; the first term is due to the induced dipole moments of all the component molecules, and the second term is due to the permanent dipole moments of the water vapor molecules. The constants are based on

experimental data.[4] The refractivity formula is accurate to 30 GHz or higher; **the atmosphere is not dispersive for radio waves.** At higher frequencies (in the infrared) the water vapor molecules can no longer keep in step with the fields, and no longer contribute to the refractivity. The refractivity for visible light is just the first or dry term.

Atmospheric pressure decreases exponentially with height. Since the dry term of the refractivity equation is dominant, the refractivity of a well-mixed atmosphere decreases almost exponentially with height. However, under certain conditions, the refractivity may deviate greatly from the smooth exponential decrease in the lower 12,000 feet of the atmosphere.

The nominal rate of decrease of temperature with height in a well-mixed atmosphere is about 2.2° C per thousand feet (3.9° F). A greater rate of decrease would promote rapid convection; a slower rate of decrease would inhibit convection. An **increase** of temperature with height is called a temperature inversion. One effect of temperature inversions is to trap water vapor and/or atmospheric pollutants. The radio refractivity may decrease rapidly within the space of a thousand feet or less through a temperature inversion that has trapped water vapor, and such a discontinuity may reflect or refract vhf and uhf radio waves.

The change of refractivity in a given distance is called the refractivity gradient. The path followed by a radio ray in the atmosphere is dependent upon its elevation angle and the refractivity gradient along the path. Usually only the vertical component is considered, and the atmosphere is assumed to be horizontally homogeneous. For rays with elevation angles below 5° or so, the curvature of the ray is equal to the negative refractivity gradient.

Thus, when the negative refractivity gradient is equal to 48 per 1000 feet or 157 per kilometer, the curvature of the ray equals that of the earth, and a ray launched parallel to the earth will follow the earth's curvature. Negative gradients of greater value are called ducting gradients. A duct is formed between the maximum and minimum heights at which a ray may be parallel to the earth—unless the minimum height is zero; in this case the ray need be horizontal only at the maximum height. These heights can be found by raytracing using Bouger's rule:

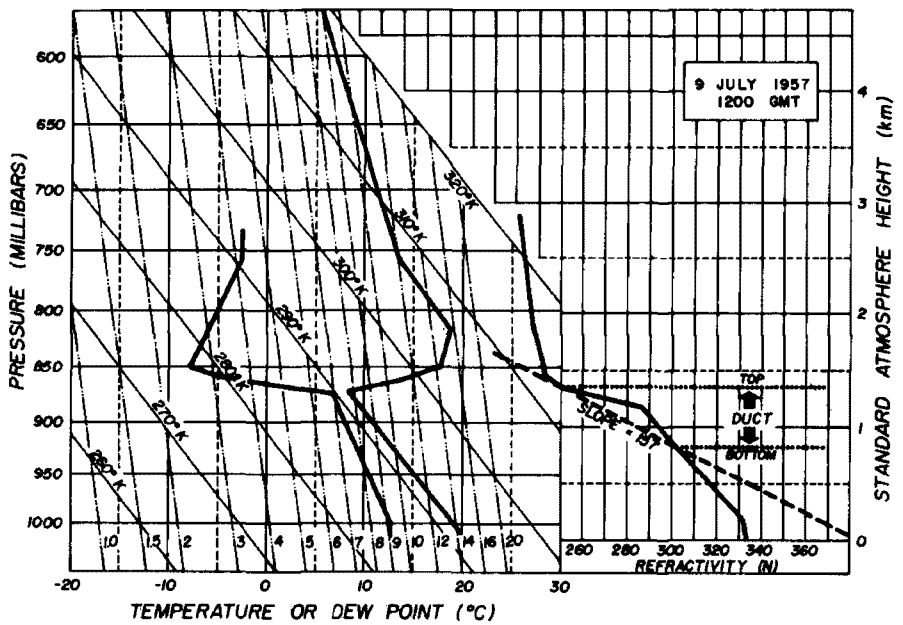


fig. 1. Upper-air sounding and derived refractivity plot for weather ship N, 1200 gmt, July 9, 1957, shows pronounced tropospheric duct. The plot on the left, water vapor, may be read as dew point or as mixing ratio in parts per thousand. The center plot is of temperature, while the plot on the right is refractivity, N. The slope of the dotted line should be — 157 per km.

$$n\ r \cos \theta = K$$

Where r is the distance from the center of the earth, θ is the elevation angle, and K is a constant for any given ray called the characteristic. (Note the difference from the formula given in the August, 1969, propagation column[3]. In that case it was elevation angle vs angle of incidence.)

There is a more simple technique for determining duct height—by graphical construction on a plot of h vs N. A line with a slope of −157 per km is constructed tangent to the h-vs-N plot at the maximum height of the ducting gradient. The line either intersects the h-vs-N plot or the ground at the bottom of the cut. In the first case it is called an elevated duct, and in the latter, a ground-based duct.

Normally a duct is formed with both a temperature inversion and a sharp lapse of water vapor content. However, if the water-vapor content remains constant, the necessary temperature gradient is 92° per km or 28° per thousand feet. On the other hand, if the temperature is held constant, the necessary drop in water-vapor content (mixing ratio) is 16.4 grams of water per kilogram of dry air per km, or 5 gm kg/thousand feet. The latter situation is the more likely.

The variation of dew point temperature and refractivity with atmospheric pressure is shown in **fig. 1.** The first two parameters were obtained from a weather bureau pseudoadiabatic* chart which was derived from a radiosonde weather balloon sounding over a weather ship midway between San Francisco and Honolulu during the first California-to-Hawaii troposheric contact on 144 MHz.

This pseudoadiabatic chart has 4 scales. The horizontal lines are the total atmospheric pressure in millibars. The vertical lines are temperature and dew point in degrees centigrade. The almost vertical lines are mixing ratio in grams of water vapor per kilogram of dry air. The lines at almost 45° are the the dry adiabats, lines of constant potential temperature are in degrees Kelvin.

* Pseudoadiabatic refers to variations in volume or pressure accompanied by loss of heat which is due to the immediate dropping out of all condensed material as soon as it is formed; used in reference to the cooling of rising air in which precipitation occurs.

At the time the chart of **fig. 1** was plotted, the surface temperature was 21° Centigrade; this corresponds to a potential temperature of about 292° Kelvin (at 1000 millibars). Free circulation to a pressure level of 876 millibars, the base of the temperature inversion, is indicated.

Note that the height scale on the right depends on surface pressure and the variation of temperature with height. The height scale plotted is for a surface pressure of 1013 mb and a **standard** atmosphere. Since the surface pressure was 1026 mb the true heights are higher than indicated.

Note that the temperature inversion extends from 1200 to 1800 meters. However, only the initial decrease in water vapor content between 1200 and 1340 meters is rapid enough to produce a ducting refractivity gradient (−252/km in this case). The extension of a line of slope −157 km from 1340 meters intersects the h-vs-N plot at 850 meters. Thus, the duct is about 490 meters thick and extends 350 meters below the temperature inversion.

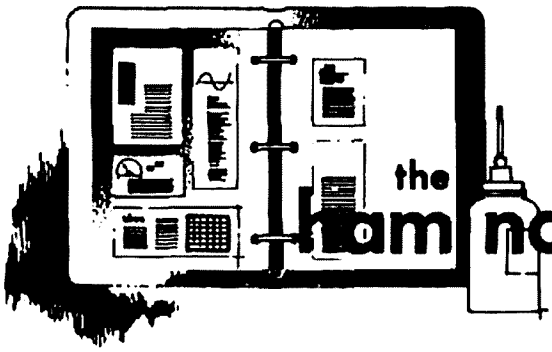
Assuming that a signal has somehow gotten trapped in the duct (perhaps the source was a low-flying aircraft inside the duct), it may travel hundreds of miles with little attenuation. Note, however, that most of the energy radiated by our hypothetical source will escape. At most, rays with elevation angles between $+\frac{1}{2}°$ and $-\frac{1}{2}°$ will be trapped.

In future articles, I will describe the frequency dependence of tropospheric duct communication, climatology of the trade-wind belts, and weather observations during some long-distance vhf communications between California and Hawaii.

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3. V. R. Frank, "Ionospheric Propagation," *ham radio*, August, 1969, p. 61.
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ham radio

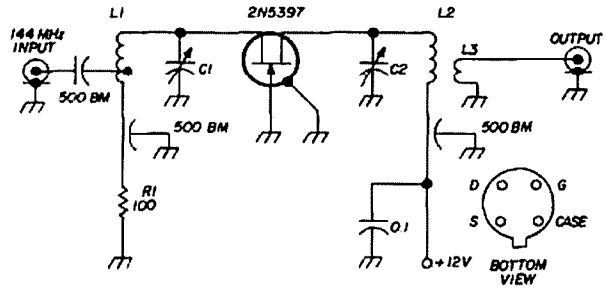


super two-meter preamp

The new Siliconix 2N5397 fet makes noise figures of 1.5 to 2 dB commonplace at 144 MHz. Although the 2N5397 is priced at $8.50 in small quantities, its performance and ease of operation make it well worth it. The 2N5397 was designed for vhf use, and it works well beyond 500 MHz, but that's another story. I was mainly interested in 144 MHz operation because a 1.5 dB noise figure is needed for moonbounce.

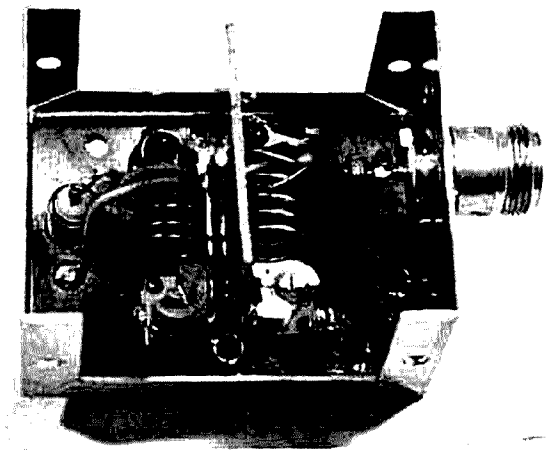
At the suggestion of the designer of the device, I tried it in a grounded gate configuration (fig. 1). Much to my delight, the circuit performed better than anything I have ever had the opportunity to use, or measure. One glance at the schematic will show you that it is as "super simple" as you can get.

Since I already have a high-performance two-meter converter, the 2N5397 preamplifier was designed to mount as close to the antenna relay as possible. This is the reason for the type-N input connector.



C1, C2 1.7-14.1 pF air variable (E. F. Johnson 189-505-5)
L1 5 turns no. 16, 3/8" diameter, 3/4" long, tapped at 1-1/2 turns from cold end
L2 4-3/4 turns no. 16, 3/8" diameter, 1/2" long
L3 2 turns no. 20 insulated around cold end of L2

fig. 1. High performance grounded-gate fet preamplifier for two meters. The 2N5397 is mounted in an IERC mounting clip type TXBE-019-021B or TXB2P-019-028B. Case and gate leads of the fet are soldered together.



I built my preamp on a 2x2 5/8-inch piece of copperclad printed-circuit board with a 1 3/4 x 1 1/2-inch shield, and then fitted into a 2 3/4 x 2 1/2 x 1 5/8-inch minibox. I prefer this type of construction because it provides good grounds.

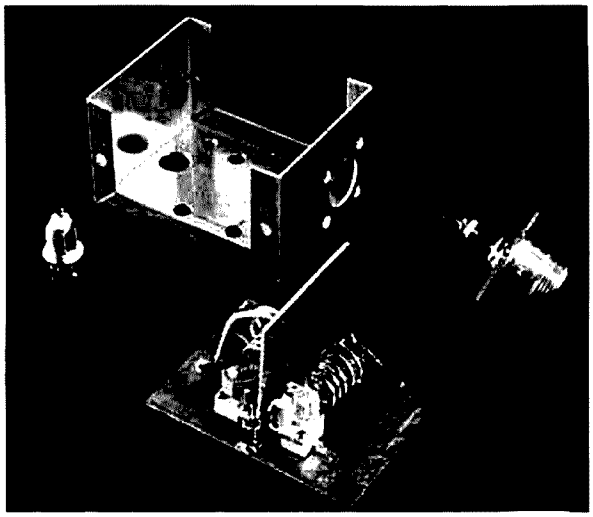
It's my opinion that the fet mounting is important to obtain optimum performance. IERC makes a TO-18 heat sink that can be soldered directly to the chassis; this is an ideal way to ground the fet case. The gate lead is soldered to the case lead to provide a good ground for the gate.

Tuneup is very simple. First apply +12

volts, and adjust R1 for 5 mA drain current. Next, feed in a 144-MHz signal and tune C1 and C2 for maximum output. Carefully adjust the position of L3 for maximum output, and readjust C2.

The gain of this fet preamplifier should be greater than 10 dB, and more likely, 15 dB. If you have a way of measuring noise figure, adjust the tap on L1 for best noise figure. If you don't have any noise-figure equipment, the tap shown in **fig. 1** will give good performance.

Ken Holladay, K6HCP



K6HCP's super two-motor preamp is built on a small piece of copper-clad board, and then mounted in a minibox.

improved sidetone operation for the SB301/401

After using my new Heath SB-301/401 combination for several weeks, I found one thing that was evidently overlooked—the sidetone monitoring system works through the speaker only, even with the earphones plugged into the phone jack. To make the sidetone monitoring compatible with earphone operation only involves the installation of a jumper from the 500-ohm **anti vox** jack to the **spare** jack diagonally across from it on the rear apron of the SB-301 receiver. The bus-wire jumper should be installed inside the chassis.

To complete the modification, it is also necessary to change the interconnecting audio cables that are used between the receiver and transmitter. The cable from the speaker is now connected to the **speaker 8 ohm** jack on the SB-301. The **speaker** jack on the SB-401 is connected to the newly wired **spare** jack on the SB-301. The **rcvr audio** jack on the SB-401 is no longer used. Other interconnecting cables are the same as before. With this simple modification, it is possible to use the sidetone monitoring system through either the earphones or the speaker. This modification should also work with the earlier SB-300/400 combination.

Don Bennecchi, W1WLZ

simplified balun

An antenna with open ends is a balanced system. Feeding it directly with coax (unbalanced feeder) causes rf radiation on the outside of the coax. The usual detuning sleeve (bazooka), a 1-to-1 transformer, requires a rather elaborate metal sleeve an electrical quarter-wavelength long. This balun has no effect on the impedance of the line or antenna. It only keeps radiation from coming back down the outer shield of the coax.

This sleeve can be considerably simplified by using kitchen aluminum foil wrapped with plastic electrical tape along its entire length (**fig. 1**). The sleeve is shorted to the coax shield at the end away from the antenna; the antenna end remains open.

Richard Mollentine, WAØKKC

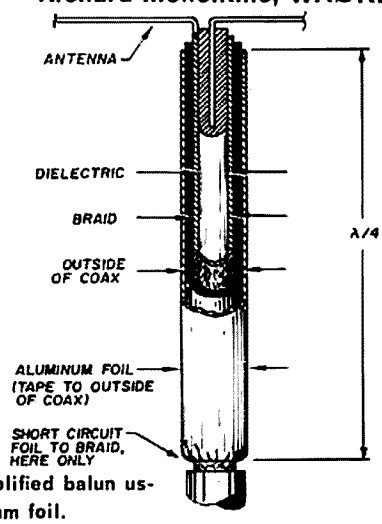


fig. 2. Simplified balun using aluminum foil.

low-loss coax

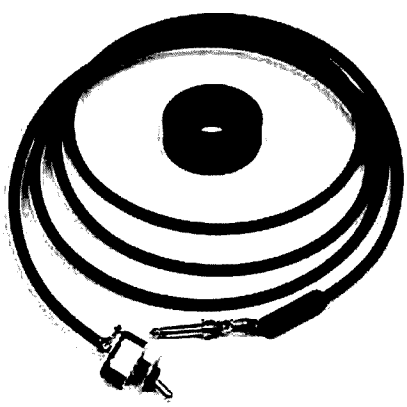
When I apparently wasn't getting out of my own backyard with a new antenna installation on two meters I was a little suspicious of that "new manufacture" RG-11/U I was using. I used a 200-watt bulb to check out the coax—when connected directly to the transmitter output connector, it was lit to full brilliance; when the coax was put between the bulb and the transmitter, the bulb barely glowed.

The solution was a section of home-made low-loss coaxial line. A 20-foot length of ½-inch rigid copper pipe was used for the out conductor; a length of aluminum clothesline wire (about no. 9) serves as the center conductor; teflon washers spaced every 8 inches keep the center conductor centered in the pipe.

Impedance is about 72 ohms and there is almost no loss. The assembly is strong enough to support small antennas (like a coaxial dipole) and is unaffected by rain; lateral support is provided by three nylon guys.

Gus Gercke, K6BIJ

a method for remote keying your transmitter



If you're tired of those endless trips up and down the tower to turn the transmitter on and off for swr checks, you'll be interested in this gadget. All you need is a two-conductor phone plug, a spare length of coax long enough to reach from antenna to transmitter (or any other two-conductor cable that's long enough), an spst toggle switch and some electrical tape. Connect the phone plug to one end of the coax, the spst switch to the other end and tape the switch terminals. Put the phone plug into the cw key jack on your transmitter and operate the switch to be sure everything is working properly.

Now load your transmitter into a good dummy load; exchange the dummy load for the transmission line to your antenna, and take the remote-keying line and swr bridge up to the antenna location. This is the procedure I use for keying my transmitter from the roof when tuning up the gamma match on my 6-meter Yagi.

Fred Hock, WA3HDU

wind-protected crank-up tower

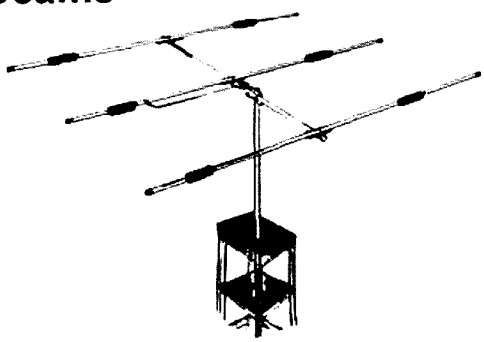
After building the anemometer described by W6GXN in the June, 1968 issue of **ham radio**, I decided to carry the project one step further. I replaced the conventional 1-mA meter movement with an instrument that has a set of contacts that may be used for external control. The contacts are adjusted so they close at some pre-determined wind velocity—this picks up a latching relay that controls the tower motor. With this setup the tower is lowered automatically when the wind exceeds a certain level.

Simpson and API both build adjustable meter relays that can be used for this application. Although these instruments are fairly expensive, you can often find surplus units with the necessary contacts to control the remote relay. When the meter needle reaches a pre-determined setting, the magnetic contacts in the meter lock the contacts together, closing the latching contacts and operating the remote relay to control the tower motor. You could even use the extra set of contacts to indicate if the tower is up or down.

E. C. Sherrill, K6JFP

new products

miniature monoband beams



The new monoband Mini-beams from Mosley Electronics, Inc. are designed to conserve both space and money. These mininature three-element beams for 10 and 15 meters use elements that are 6 to 8 feet shorter than their full-sized counterparts; furthermore, they weigh about one-third less. The MB-10 Mini-Beam for 10 meters weighs about 6½ pounds; the MB-15 for 21 MHz weighs 9 pounds. Because their lighter wind loading puts less strain on tower and rotor, a less expensive rotor can be used. The new Mini-Beams feature a gamma matching system, time-tested Mosley boom-to-mast clamping arrangement, high grade aluminum tubing, high impact polystyrene clamping blocks and stainless steel hardware. Rated for 500 watts cw or 1 kW PEP ssb. For more information, write to Mosley Electronics, Inc., 4610 North Lindbergh Boulevard, Bridgeton, Missouri 63042.

six-meter linear



The Raytrack Company has just introduced the Horizon VI L, a 2000-watt PEP linear amplifier for operation on the amateur 50-MHz band. The amplifier uses two zero-bias grounded-grid 3-500Z triodes, operated in class B for maximum efficiency. The linear uses a separate heavy-duty solid-state power supply, an adjustable alc network that is compatible with most standard equipment, a relative output indicator, and exciter feedthrough provided by the flip of a switch—no need for additional switching when you want to operate with just the exciter. When you want to put the linear on, the fast-heating 3-500A's are ready to go in a minimum amount of time. The Horizon VI L is easily driven to full output by 100-watt PEP six-meter transmitters. Unit can also be used on cw, a-m and RTTY. $595 including power supply from your local dealer, or write to Raytrack Company, 3498 East Fulton Street, Columbus, Ohio 43227.

experimenters bulletin

If you're a typical basement electronics experimenter, you'd probably enjoy reading **The Australian EEB,** an informal electronic experimenters bulletin published about eight times a year. Edited by VK7RG and VK7ZAR, it is an experimenter's magazine, but not necessarily a construction manual—it contains many constructional projects, but also many common sense observations on technique. It is a magazine which ranges over a wide field of interests in electronics, and

provides a voice for experimenters who have ideas, and who want to tell others about them informally, by letter or by article. The style is very casual, and the format haphazard but easy to read and enjoy. Topics covered in the past include diode and transistor testing, meter protection, printed circuit techniques, transistorized voltmeters, Q-multipliers, oscillators and a complete series on transistorized transmitters. Current issues are loaded with common sense ideas for receiver and mixer stages. Subscriptions are $1.80 per year and well worth it. Write to The Australian EEB, P. O. Box 177, Sandy Bay, Tasmania 7005, Australia. Sample copies are available.

rf power transistors



Motorola Semiconductor is currently developing a series of transistors and integrated circuits for the microwave region. The first devices to be made available are characterized as oscillators and amplifiers. Oscillator transistors in the series are the MM8008, 8010 and 8011. The MM8008 develops a minimum of 300 mW at 2 GHz and more than ½ watt at 1.5 GHz. The MM8010 and 8011 have somewhat lower power outputs.

These transistors are housed in the wide-flange TO-107 case shown in the photo. This package provides both convenient mounting and efficient heat transfer in resonant cavities and is less costly than conventional coaxial packages. A specific feature of these oscillator transistors is their low noise generation; in oscillator service, a signal-to-noise ratio of better than 80 dB can be obtained.

The second series of rf transistors is designed for class-C service in the 1-GHz range. The MM4430 is capable of 2.5 watts output at 1 GHz with 6 dB gain; its 1-W counterpart is the MM4429. These transistors are packaged in a ceramic package with low inductance stripline leads. The stripline opposed emitter (SOE) package has emitter leads on opposite sides of the package. This provides isolation between the base and collector leads and permits low inductance grounding.

Two other transistors which may be used as either oscillators or amplifiers in the 1 GHz region are the 2N5108 with 1 watt output and 5 dB gain at 1 GHz, and the MM8009 with slightly less output.

Informed sources at Motorola indicate that rf transistors due to be introduced by late this year include power transistors with 5-watt capability up to 2 GHz and 10 watts at 1 GHz. For more information on these new microwave power transistors, write to Technical Information Center, Motorola Semiconductor Products, Inc., Box 20924, Phoenix, Arizona 85036.

complementary audio transistors

Motorola has just announced two pairs of low-cost plastic encapsulated silicon transistors for use in 20- and 35-watt complementary audio amplifiers. The four transistors use Motorola's Thermopad package for easy mounting and efficient heat transfer. The 5-amp npn MJE205 and pnp MJE105 will dissipate up to 65 watts, and exhibit current gain of 25 to 100 at a collector current of 2 A. An application note describing their use, "A 20-Watt Amplifier with Complementary Symmetry Thermopad Silicon Power Transistors," AN433, is available at no charge.

The npn MJE2801 and pnp MJE2901 are 10-amp transistors with 90-watt power dissipation and high current gain of 25 to 100 at a collector current of 3 amps. An application note describing their use in 35-watt amplifiers is available (AN427). For more information, write to the Technical Information Center, Motorola Semiconductor Products Inc., Box 20924, Phoenix, Arizona 85036.



comments

Dear HR:

Concerning the July **ham radio** editorial on antennas, my only worry is for sincere experimenters who want to try Bonadio antennas. They may be discouraged by knowledgeable engineers who "know" that these antennas are only, "normal antennas—nothing special about them. Any expert engineer can see that; he does not have to try them."

Actually, the last great frontier in ionosphere propagation is in the approximate 88 dB between excellent skip and the unreadable level. Engineers have insisted to me, "A wave is a wave, is a wave, and they all propagate alike." However, we know that polarized light reflects at different rates from different surfaces, and in different angles, and in different densities in propagation, and when detected after reflection and refraction the light-wave losses can vary by 50 dB from maximum to minimum.

My experiments suggest to me that waves from my antennas, in overseas contacts, propagate with less losses only when the path losses are significant. I interpolate differences of +10 dB commonly, +20 dB frequently, +30 dB occasionally and once in awhile, apparently +40 dB. However, when ionospheric skip is wide open, my antenna is only equal to a Hertz antenna which is, then, very satisfactory.

The last two DX cards I received were from PY1DMS, in Rio, on 40 meters, who gave me S9+5 while I never could read him to hear my report, (about 45 dB between us) and from PY1DEH, in Guanabara, on 15 meters. He gave me S9+10 but never was better than readability 2, and I had to wait for the card to find a report (about 50 dB between us).

This is a problem. While they hear me *well*, I have *all I can do to hear them* over the cosmic noises. Its a little like being on the same side of one-way skip most of the time.

Meanwhile, every one of 157 antenna manufacturers have turned me down, yet none experimented with the ionosphere in the circuits under test. This is like my 15 months in Africa where I avoided the locations of elephants, being proof to me that there are no elephants in Africa. The excellent performance characteristics of my antennas are particularly apparent when the ionosphere is the most lossy, so tests without including the ionosphere can not be valid.

I am pleased to think that amateurs can enjoy these antennas alone for years before the commercial interests admit that amateurs developed and proved out something that the commercials had overlooked. We need these occasions to justify our occupying the spectrum.

George A. H. Bonadio, W2WLR



"If you'd use a 20,000-ohm resistor instead of 10,000 ohms, the 1 mA change in plate current will develop 20 volts, giving you a nice increase in amplification."

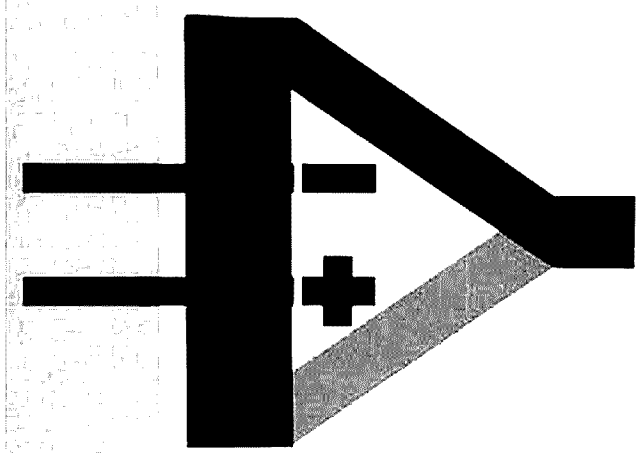
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ham radio magazine

NOVEMBER 1969

OPERATIONAL AMPLIFIERS...



THEORY, SELECTION, & APPLICATION

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staff

editor
James R. Fisk, W1DTY

vhf editor
Nicholas D. Skeer, K1PSR

fm editor
J. Jay O'Brien, W6GDO

associate editors
A. Norman Into, Jr., W1CCZ
Alfred Wilson, W6NIF
James A. Harvey, WA6IAK

art director
Jean Frey

publisher
T. H. Tenney, Jr. W1NLB

offices

Greenville, New Hampshire 03048
Telephone: 603-878-1441

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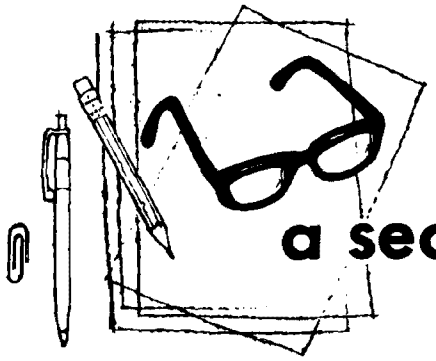
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a second look

by jim fisk

Spring and early summer seems to be the traditional time to install a new antenna (for most sensible people—I always seem to wait until winter's first real blizzard). Magazines often feature antenna articles in the spring, and it's great to get outside and try out new ideas. In the fall, the new antenna is given a final inspection and maintenance is performed to ensure that it will withstand the rigors of coming winter storms.

Regardless of when you work on an antenna installation, a very real physical risk is involved. When you climb a tower or a steeply pitched roof, one incautious moment can end in tragedy. With winter just around the corner, many amateurs will be thinking about winterizing their antennas. I'd like to offer a word of advice: observe safety rules and use the buddy system, even though you may have to climb only fifteen feet.

Here's a true story about a ham in California who was making antenna adjustments alone. The guiding forces that protect drunks and other foolish creatures must have been operating in this instance, because this fellow lucked out. Others may not be as fortunate.

Bill G. (not his real name, of course) decided to work on his antenna one afternoon. Coaxial cable needed replacing and some soldering was necessary. Bill started up a ladder laden with 70 feet of coax and a 50-foot extension cord looped around one shoulder. Over this was looped a leather holster full of tools. A bag of hardware was tied to the bottom of the tool holster. Bill climbed over the end of the ladder and started up the steep roof. No problem here: he'd done this many times before. But he didn't account for one important factor. A typical Southern California breeze, about 15 knots, was blowing in from the ocean.

When Bill reached the base of the antenna mast, he was standing on the peak of the roof, which offered very little footing. You can probably guess what happened next. The weight of gear and tools on one side of his body, plus the wind forces, caused him to lose his balance. Friend Bill, tools, cable, and hardware started down the roof toward the ground 30 feet below. Somehow the coax cable wrapped itself around one of the standpipes on the roof, and Bill ended up dangling over the edge of the eves. Thanks to the strength of the cable and short overhang of the eaves, Bill managed to reach the ground by going down the side of the house rappel fashion. (Antenna work was suspended for the rest of the day.)

It is easy to overlook safety precautions when you're preoccupied with the problem at hand, when you're in a hurry, or if you're tired. Climbing a tower or roof should never be attempted without another person standing by to give aid or summon help. Professional tv antenna installers, for example, never work alone. Insurance companies and common sense forbid it.

Overconfidence when working on high structures is probably the greatest cause of accidents. Our California friend had been on his roof many times during the weeks preceding the accident. He felt almost as much at home working on his roof as in his workshop. The first time up, he took every possible precaution. But as he became familiar with the situation, he became careless. Result: near disaster.

The consequences of a 30-foot fall can be just as permanent as a jolt from a 3-kV power supply. When you work on your antenna this fall, remember the tale about our colleague who almost ended up a statistic. Be careful, and use the buddy system.

Jim Fisk, W1DTY
editor

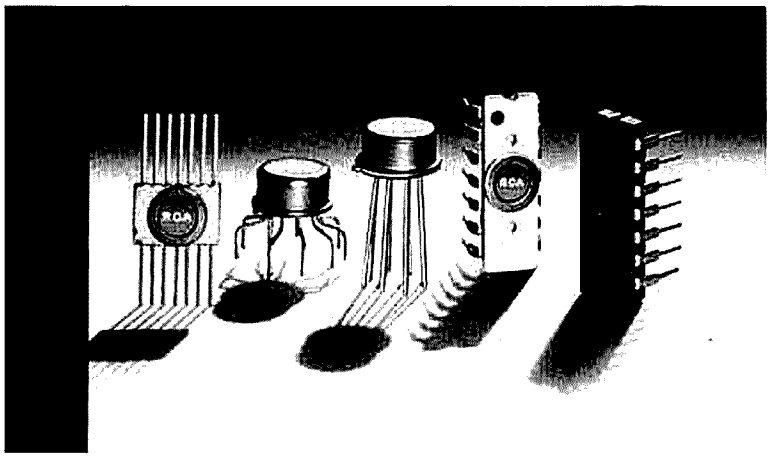


photo courtesy RCA

what's this we hear about op amps?

The operational amplifier has invaded amateur radio—here are the basic facts on its theory, selection and application

Don Nelson, WB2EGZ, 9 Green Ridge Road, Ashland, New Jersey 08034

If you have been keeping up with the current electronics literature, you've surely seen articles on the integrated circuit operational amplifier. Perhaps you've wondered just what it is and what it can do for you. The fact is, it can do just about anything you wish, and do it better than conventional circuits.

Tube versions of the op amp have been around for a long time. They were originally used in analog computers to perform mathematical operations such as addition, subtraction, and averaging. The main objection to these circuits was their huge physical size. Recent advances in solid-state technology have produced op amps, at very reasonable cost, with active elements formed on a single chip of silicon. A complete amplifier now occupies less space than many of the discrete components used in the original tube circuits.

The IC op amp is so useful in amateur radio applications that I've prepared this article to acquaint you with it. The first part of the article discusses some of the more popular circuits and gives the equations describing the relationship between input and

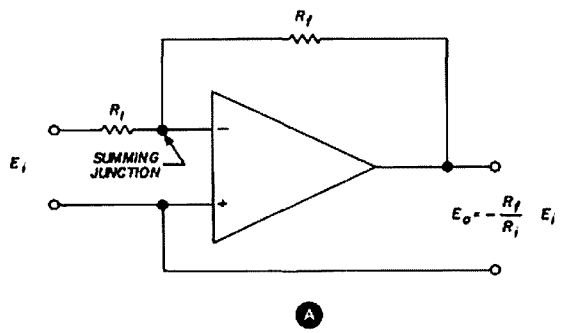
output. Then comes a description of the op amp's gain characteristics. The last part of the article is devoted to some applications you'll find useful around the amateur station.

typical circuit

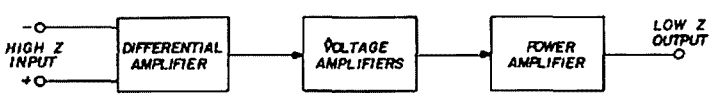
The input stage of the op amp is a high-impedance differential voltage amplifier. This is followed by other voltage amplifiers. The output stage is a low-impedance power amplifier.

Fig. 1 shows a typical circuit. Resistor R_f feeds the output to the negative input, which is sometimes called the summing junction. The negative input is isolated from the driving signal, E_i, by resistor R_i, which represents the circuit's input resistance. The negative input is 180 degrees out of phase with the output and is at ground potential. Under these conditions, no current flows into the amplifier, because the current in R_f and R_i is equal and opposite. Ohm's law says that the output voltage, E_o, is related to the input voltage, E_i, in the same proportion as the values of R_f and R_i. The negative sign in the equation of **fig. 1** means that the phase has shifted 180 degrees.

fig. 1. Typical operational amplifier circuit, A, and block diagram of its three stages, B. Many arrangements of feedback elements are possible.



A



B

definitive examples

The following op amp circuits are ideal representations. Nothing is perfect, of course, but I've used examples of a perfect amplifier to provide definitive examples. A perfect amplifier would have these characteristics:

1. Infinite open-loop gain.
2. Infinite input impedance and zero output impedance.
3. Zero response time (the output changes simultaneously with changes in input).
4. Zero offset. With no voltage between the input terminals, the output voltage will be zero.

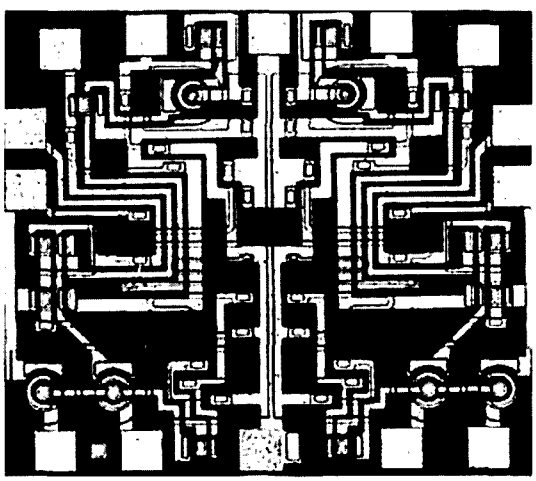


photo courtesy Motorola

The integrated-circuit chip used in the Motorola MC1535 dual operational amplifier.

The important things to remember about these characteristics, which are called summing junction restraints, are:

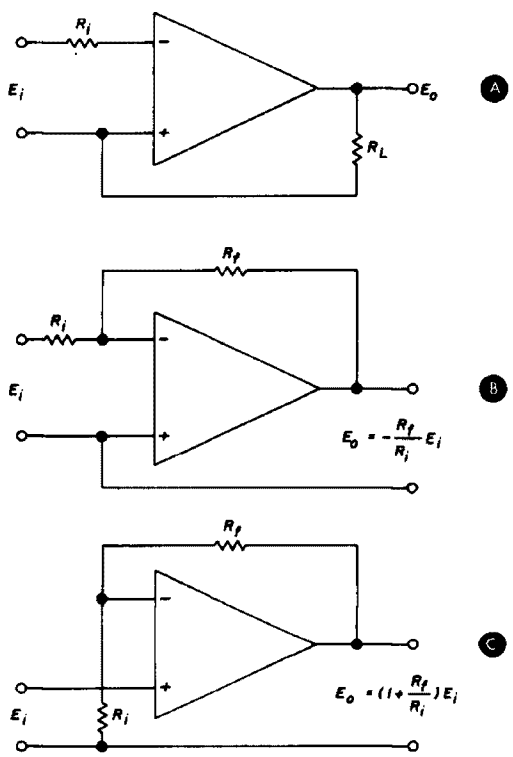
1. No current flows at either positive or negative input.
2. Both inputs are at the same potential.

open-loop operation

No feedback is used in the circuit of **fig. 2A.** The amplifier is running wide open. If the input is other than zero, the amplifier will be driven into saturation. It isn't often

used in the open-loop mode because of practical considerations. One use, however, is in a voltage comparator circuit. If two ac voltages are applied to the input, the open-loop amplifier will follow their potential difference. As the voltage on the positive terminal changes from that on the negative terminal and vice versa, the amplifier will swing as far as its supply will allow.

fig. 2. Three common op amp circuits. An open-loop circuit is shown in A, useful as a voltage comparator. B and C are inverting and non-inverting amplifiers.



inverting and noninverting amplifiers

Two widely used arrangements of the op amp are illustrated in **figs. 2B** and **2C.** In the circuit of **fig. 2B** (which is the same as that of **fig. 1**), the output signal is inverted with respect to the input. If a square wave, for example, with positive-going pulses is applied to the input, the output pulses will be negative. Gain will be proportional to R_f and R_i. The sign in the right-hand member of the transfer function (relationship of input to output) will be negative because of the 180-degree phase reversal.

In the noninverting circuit of **fig. 2C**, input and output are in phase, which accounts for the plus sign in the equation.

voltage follower

A variation of the noninverting amplifier, the voltage follower, is shown in **fig. 3A.** Note that the output is connected to the negative input. The positive input is driven directly by the input signal, E_i. Output is equal to input: a unity-gain amplifier. This circuit is used for following voltage references. The limitations of the cathode follower (or emitter follower in transistor circuits) are minimized.

transducer, adder, subtractor

The circuit of **fig. 3B** is a current-to-voltage transducer. It can be used to drive a meter, recorder, or other voltage-operated indicating instrument from limited current sources.

Voltage inputs are added directly in the summing amplifier of **fig. 3C.** The op amp is shown here in one of the operations for which it was originally used. Each input may be weighted by using different resistor values. Input weighting is proportional to the gain of the particular input: E_1 will have a weight of 2 if $R_f = 2k$ and $R_1 = 1k$. If $R_2 = 500$ ohms, E_2 will have a weight of 4.

The circuit of **fig. 3D** is sometimes called a balanced input amplifier or symmetrical subtractor (difference amplifier). It's used when neither side of the signal being amplified is at ground potential, as across a current-sensing resistor. Other inputs may be added where inputs to the negative terminal are additive, and those to the positive terminal are subtractive.

integrator and differentiator

By using a capacitor in the feedback loop (**fig. 4A**), the op amp may be used to integrate voltage waveforms. When the capacitor is in the input, **fig. 4B**, the signal is differentiated. Both differentiator and integrator, as shown, are purely theoretical.

practical limitations

Most errors in a practical amplifier with known characteristics can be calculated. If

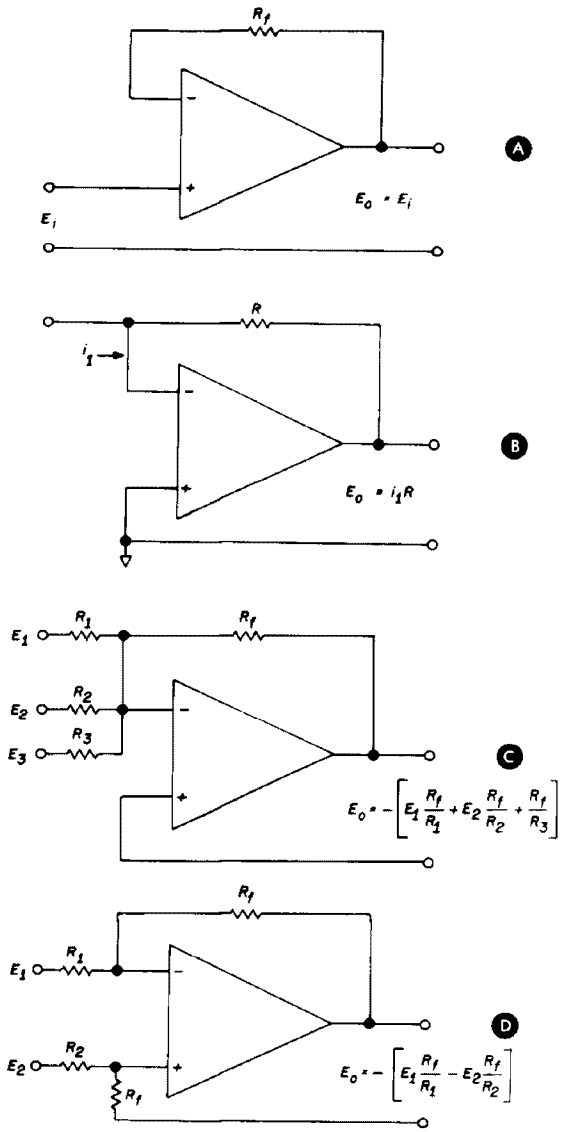


fig. 3. Other examples of op amp voltage amplifiers. A voltage follower is shown in A; current-to-voltage transducer, summing amplifier, and difference amplifier are shown in B, C and D.

the amplifier is properly chosen for a particular application, these errors may be negligible or can be compensated. With an understanding of amplifier gain, frequency response, and phase shift, you'll be able to apply compensation methods to tame the op amp of your choosing.

gain

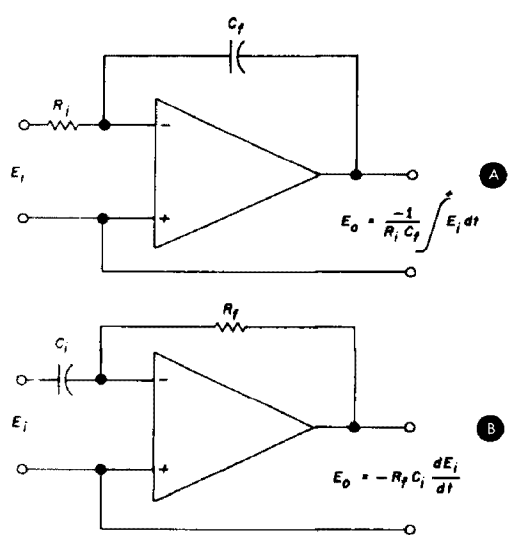
Several definitions of gain must be understood:

1. **Open-loop gain.** This is the ratio of output-to-input voltage at any frequency. No feed-back is used. Typical open-loop gains are from 10^4 to 10^9 in commercially available amplifiers.
2. **Closed-loop gain.** When feedback is used, the amplification is called closed-loop gain. For reasons to be discussed, closed-loop gain is rarely less then unity.
3. **Loop gain.** This is the difference between open and closed-loop gain. Usually, errors are minimized with greater loop gain.

Characteristics such as gain attenuation with frequency (also called roll-off) and phase shift, which are common to all amplifiers, are especially important when considering operational amplifiers. As mentioned previously, phase shift through the amplifier must be 180 degrees when feedback is employed. Any additional phase shift must be compensated, or the circuit will oscillate.

In fig. 5, gain-bandwidth characteristics are shown for an uncompensated amplifier (not necessarily typical). The phase shift (lag) increases as the gain is affected by feedback. The amplifier becomes unstable when the roll-off exceeds 18 dB/octave because of the 180-degree phase lag. In well-designed amplifiers, this limit occurs below unity gain. Even with compensation, the

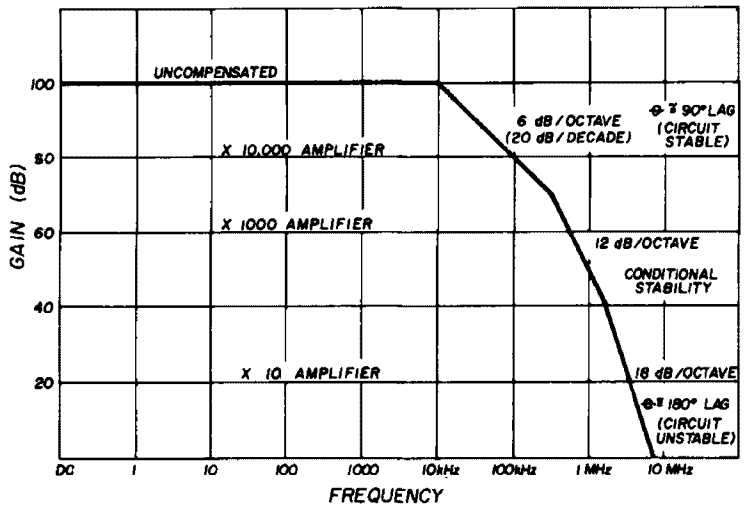
fig. 4. Voltage waveform integration and differentiation may be performed by A and B. These circuits are used for precise filtering.



amplifier can't be controlled when 18 dB/octave is reached; therefore, operating below unity gain is usually impractical. Some amplifiers may be difficult to control at gains slightly above unity.

Output limiting is another popular form of compensation (**fig. 8**). Amplifiers such as the Fairchild μA709 have a special terminal just for this purpose (**fig. 8A**). The technique of **fig. 8B** is also useful. Output compensation frequently is used to supplement some form of input compensation such as that suggested in **fig. 8A.** While every compensation problem is unique, we may generalize and say that the compensations shown above are required by the peculiarities of integrated circuits.

fig. 5. Gain-bandwidth characteristics of uncompensated amplifier. Instability occurs as gain attenuation exceeds 18 dB/octave because of 180° phase lag.



compensation

Amplifier compensation will limit frequency response, but roll-off and phase shift will be controlled. A plot of a compensated amplifier's response is shown in **fig. 6.** This type of presentation is called a Bode plot; it illustrates the limited gain roll-off (rate of closure with the unity-gain point in the frequency response of the amplifier).

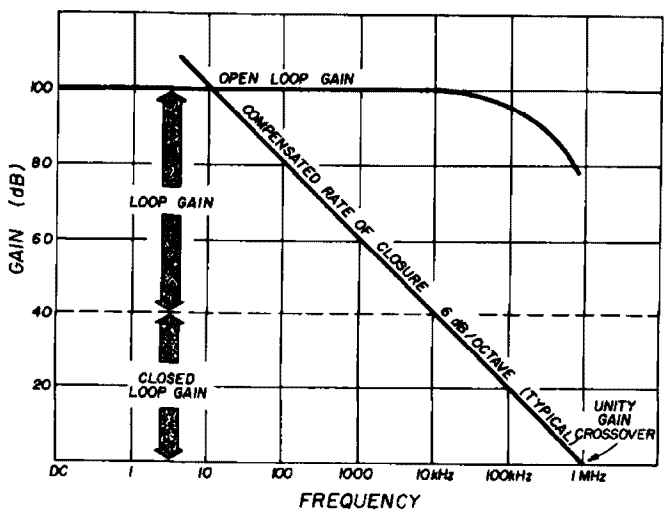
Most op amp data sheets give enough information to make a Bode plot. This will allow you to analyze the results of intended compensation. The Bode plot is the easiest way of showing the characteristics of compensation.

The simplest way to stabilize an amplifier in which a large amount of feedback is used is to bypass some signal point in the circuit. IC op amps such as the RCA CA3047 and Fairchild μA702 have terminals specifically provided for this. **Fig. 7** shows an example. If the bypass capacitors (**fig. 7A**) are increased in value, the amplifier open-loop response will shift to the left (**fig. 7B**).

As the Bode plot shows, the high-gain, high-frequency characteristics are very limited with this configuration. The simple addition of a series resistor with the bypass capacitor will yield greater bandwidth.

Amplifiers using discrete components are internally compensated to a degree. With discretes, the main compensation is for output loading, C_L, stray wiring capacitance,

fig. 6. Bode plot of compensated amplifier. Response is limited so that 180° phase shift occurs before unity gain is reached.



and input loading, C_I. Compensation techniques are shown in **fig. 9.** This is by no means the last word in compensation; it's only intended to help you when some published circuit won't work.

offset error

Among the imperfections of a practical op amp is the mismatch of components that prevents the amplifier from having exactly zero output with zero input. This may well be the most serious problem you'll encounter in dc operation of a high-gain amplifier. The basic compensation methods are shown in **fig. 10.**

input/output limitations

Input impedance and voltage swing generally may be neglected in the conventional inverting amplifier shown in **fig. 10.** The input impedance will be equal to the input resistor, R_i, because the input is a virtual ground. The amount of drift may be considered a limitation of input. The simplest compensation for this is shown in **fig. 10B**, where a resistor is used in the positive input.

Drift may be compensated similarly in a noninverting amplifier as shown in **fig. 11.** The difference here is that R_c is now the input resistor. In this circuit, the input voltage swing is a limitation. This is called "common-mode voltage swing" on the data sheet.

Output impedance, being some value greater than zero, will introduce small circuit errors. It is desirable to keep Z_{out} low.

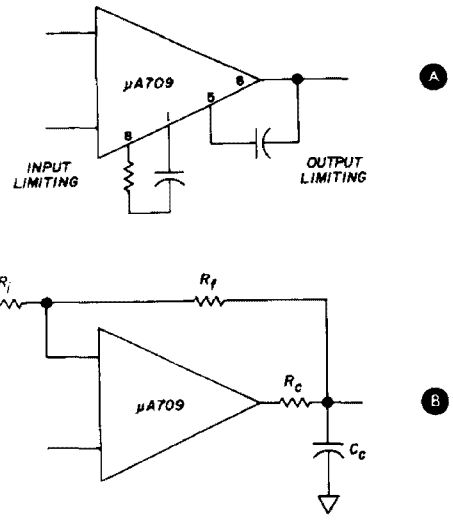


fig. 8. Response limiting at amplifier output. Capacitive compensation, A, or the RC network in B are frequently used to supplement input compensation.

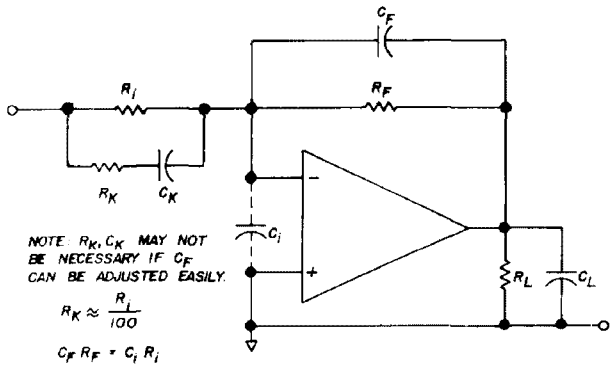
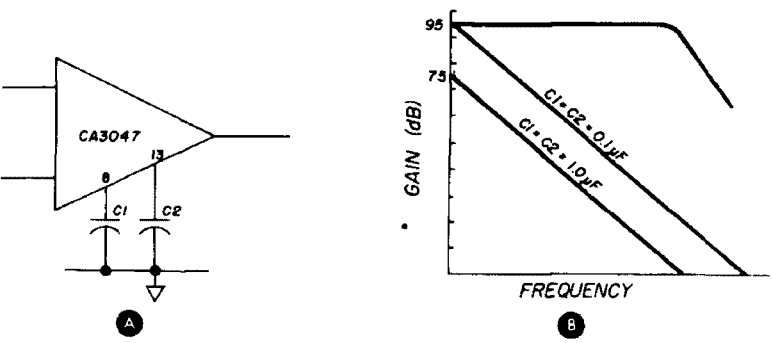


fig. 9. Compensation for op amps with discrete components. Input loading, stray wiring capacitance, and output loading must be compensated.

fig. 7. Compensating by bypassing. As capacitor values increase, A, amplifier open-loop response moves to the left, B.



This may be done by using the greatest loop gain possible. A booster amplifier also reduces Z_{out} although such an addition probably wouldn't be considered unless the output current capability is too small. The most common current booster is the pass transistor in a precision power supply as suggested in **fig. 12.**

Beware of the limitation of output voltage swing, especially in IC amplifiers. This is a luxury that closely relates to the price of the op amp. The 30 V p-p capability of RCA's CA3047A is high for an IC. In a discrete-component amplifier such as Fairchild's A00-7, the output capability zooms to 200

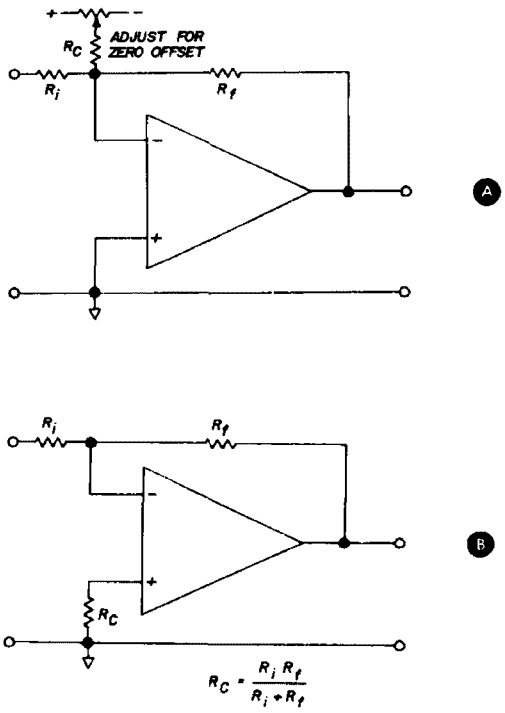


fig. 10. Compensation for dc offset error in an inverting amplifier. Adjustable current offset compensation is shown in A; drift-compensating resistor in B.

V p-p. Unfortunately, the price increases by 40 times.

One input error that's elusive—at least to me—is the common-mode error. Variously defined, this error arises from the effect of a change in one input on the signal fed to the other input. Common-mode error is smallest when the common-mode rejection ratio is high. This error is important when differential inputs are used, or when the amplifier is operated in the noninverting mode.*

selection hints

Always look for the least-expensive amplifier that will satisfy your requirements.

*Slewing rate is another limitation of practical op amps. Briefly, it is the maximum rate of change of output voltage with time. It must be considered when pulses of fast rise time are employed. It also is a limitation to using operational amplifiers at high frequencies. A thorough treatment of this parameter would be quite lengthy; however, a careful examination of the Bode plot will show how slewing rate will change with compensation. Interested readers will find a discussion related to an integrated-circuit op amp (MC-1530) in the Motorola *Integrated Circuit Handbook*, 1968 edition, p. 10-74. **Editor**

Some suggestions to guide the newcomer are outlined below.

1. High loop gain is desirable. Usually this implies the need for high open-loop gain.
2. Sufficient output voltage swing and output current to the load must be considered.
3. Offset voltage and drift must be checked for compatibility with your circuit.
4. Offset current is particularly important in circuits such as the current-to-voltage converter.
5. Common-mode voltage is important for noninverting and dual-input circuits.
6. Power-supply ripple, drift, and regulation are most important when the supply is used as a reference. However, all op amps work better with a high-quality supply.

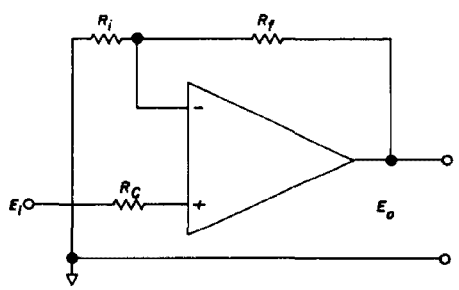


fig. 11. Drift compensation in a non-inverting amplifier. Input voltage swing is a limitation; Rc is now the input resistor.

The best over-all performance in op amps is obtained from those using discrete components—in fact, tube types. The least expensive and most interesting to experimenters are the integrated-circuit op amps. Despite their low cost, performance is quite good.

construction and layout

If you understand the parts of this article dealing with the ideal operational amplifier and the limitations of practical circuits, you're almost ready to warm up the solder-

ing iron. First, however, I'd like to give a few precautions on layout and choice of components.

capacitors and bypassing

Poor layout in an op amp circuit may cause its response to peak at the higher frequencies. Under certain conditions, oscillation will result. The problem can exist even with a neat layout. In stubborn cases, peaking may be cured with a mica bypass capacitor (try 100 pF) directly at the noninverting input. This is appropriate only for an inverting amplifier. The problem is rare when the amplifier is used in the noninverting mode.

More frequently, oscillation results from improper bypassing in the power supplies. A 0.1-μF capacitor on each power-supply lead at the amplifier socket is good practice. Low-inductance, laminated ceramic capacitors are perfect for this.

Capacitors can be critical in some circuits where low leakage is important. Dura-mica types are excellent for compensation purposes. High values and higher precision,

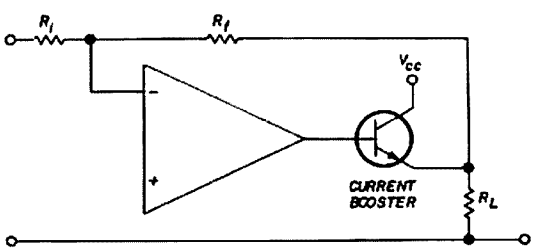


fig. 12. Increasing current output. If output current capability is too low, a booster amplifier can be used to reduce output impedance.

such as would be required for timing circuits, will call for Mylar or Polystyrene capacitors.

resistors and diodes

The giant called **loop gain,** which is restrained by an operational system, will create problems when noise and unwanted reactances exist. Therefore, certain precautions must be observed with respect to other circuit components.

Resistors must be chosen with care in systems where accuracy depends on the resistor. Wirewound resistors have low noise and excellent stability. However, they have the largest shunt capacitance and series inductance of all types. Also they're not usually available in values above one megohm, and they're expensive.

Carbon composition resistors shouldn't be

Photomicrograph of the RCA CA3033 operational amplifier. The CA3033 is the dual-in-line ceramic package version of the CA3047.

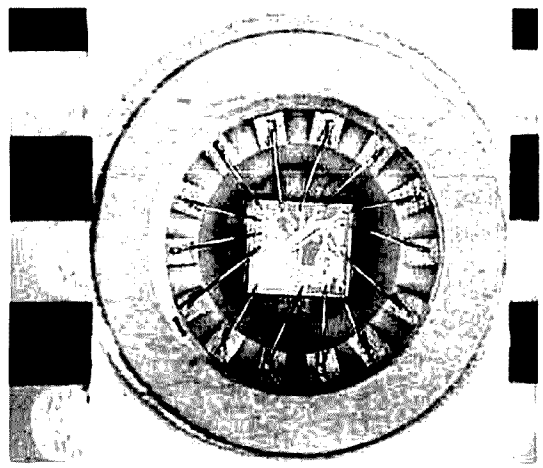


photo courtesy RCA

used where high stability is required, such as in the input and feedback circuits. Although they produce noise, these resistors are inexpensive and are satisfactory in less critical parts of the circuit.

Metal film resistors have excellent characteristics and provide a good compromise between the wirewound and composition types. Their upper range is ten megohms. Higher resistance values are available from Victoreen and Pyrofilm in the form of glass-enclosed, deposited-carbon construction. While there's little choice in precision resistors above ten megohms, you should be aware that some high-resistance types are voltage sensitive. They're not precise at voltages other than their test voltage—usually 10 V. Be careful not to get dirt or perspiration on these, as it may reduce their resistance.

Clamping diodes are frequently used in op amp circuits. Low-leakage, low-capacitance diodes such as the 1N914 should be used. **Never** use germanium diodes!

triple grounds

Three grounds should be used: **signal ground, power-supply ground,** and **chassis ground.** This triple grounding technique is essential to minimize voltage drops that would create system errors. At some point, all grounds may be connected, but not necessarily. Consider each system with respect to the voltage drops that will develop. For example, with high output current (load current), the load return to power supply ground must be direct. The reference signal, using signal ground, must not be transmitted through the same wire. **Fig. 13** illustrates some basic grounding techniques; however, the subtleties of the ground loop aren't always easily controlled. A little experimentation with the preceding concepts in mind could lead to a better solution.

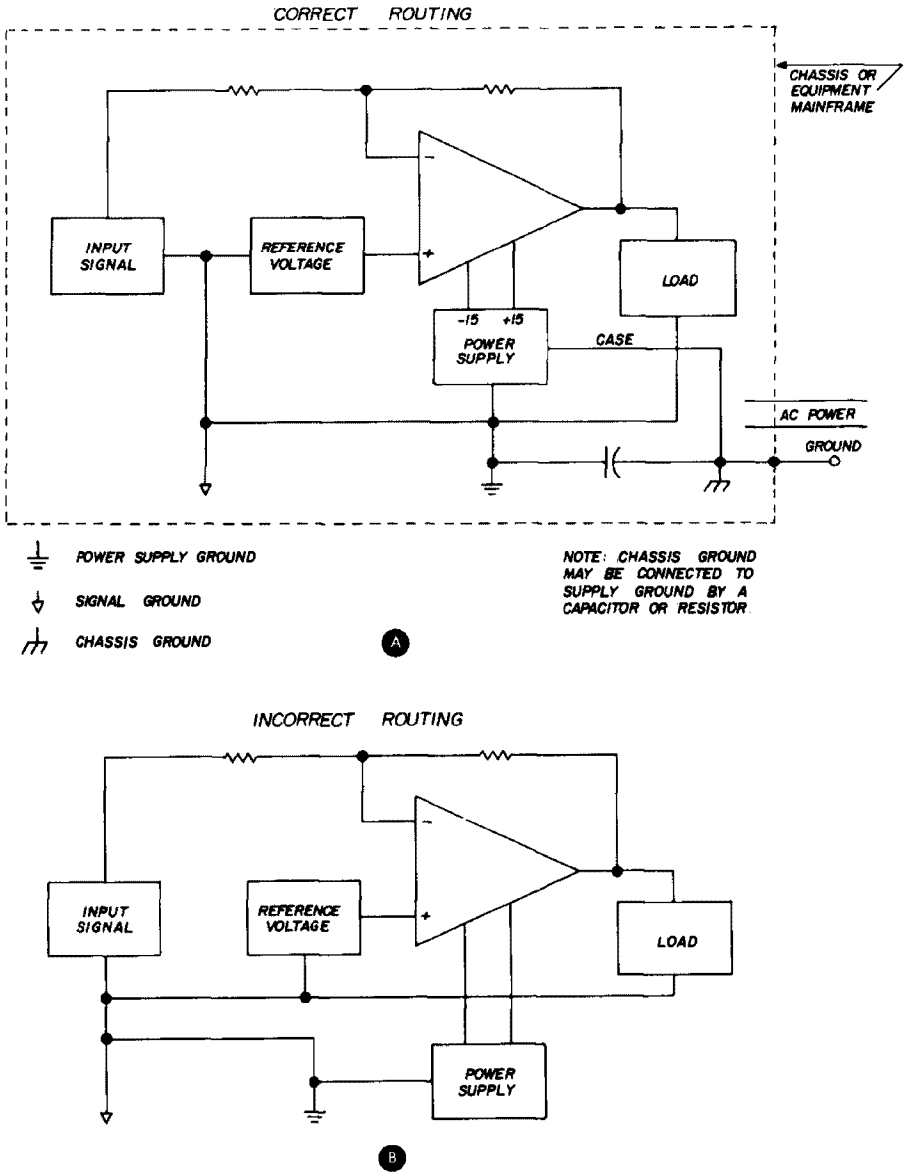


fig. 13. Correct routing, A, and incorrect routing, B, of load return in an op amp layout. Incorrect routing will cause an error in reference voltage.

a compendium of op amp circuits

I've devoted the remainder of this article to a description of some of the more common applications of the operational amplifier.

These circuits are just a starting point. I'm sure that ham ingenuity will result in many more interesting variations. Who knows? Perhaps someone will adapt one of these circuits to a communications problem and revolutionize the industry. In any event, I hope these ideas will inspire more experimentation. If you come up with a new use for the op amp, the market is wide open for your ideas.

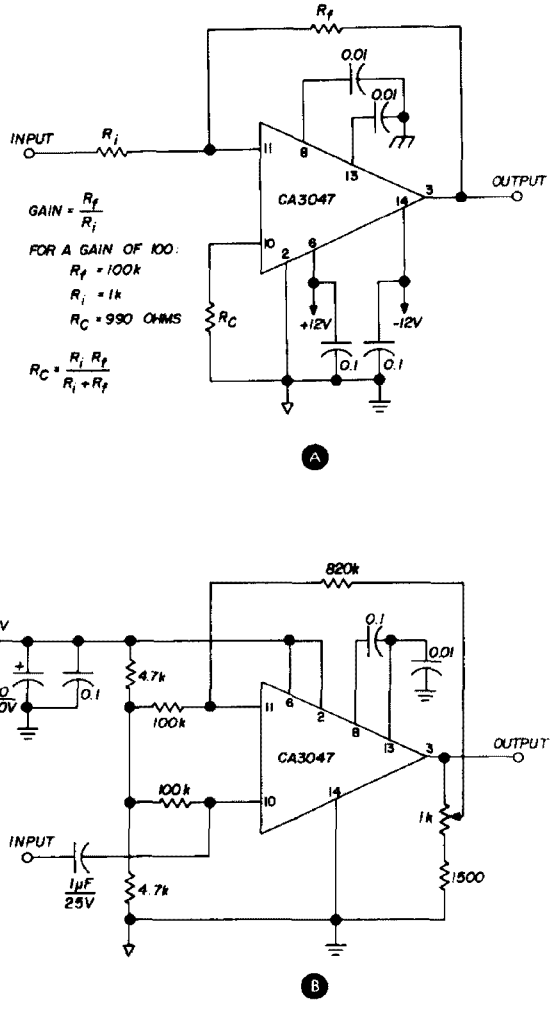


fig. 14. Complete circuits for the basic inverting amplifier, A, and non-inverting amplifier, B. Gain of noninverting circuit is approximately 10.

basic computer circuits

While basic computing circuits may not be your idea of a construction project, such applications of the op amp serve to identify what follows. As a matter of fact, with a little thought and planning, these circuits might be just the thing for a science fair presentation.

To recap, the basic inverting and noninverting op amps are shown in **fig. 14** with all the component values. You'll recall that the inverting amplifier shifts the phase of the input signal 180 degrees; that is, a positive-going input produces a negative-going output. The output signal will be in phase with the input in the noninverting amplifier.

Typical compensation is shown in the circuits of **fig. 14.** The following circuits are simplified. Compensation and proper bypassing are essential, of course. The RCA CA3047 is inexpensive and altogether adequate for the applications shown.

An adder is shown in **fig. 15.** The offset network is typical for all computing circuits. An alternate would be a voltage offset circuit, which is usually connected to the positive input. The currents from these three inputs are summed, and the negative of this sum appears at the output. Feedback at the negative input means that the input is a virtual ground, so the three inputs are effectively isolated, and no interaction exists between them.

An adder-subtractor circuit is shown in **fig. 16.** Note the equation of the circuit: the output voltage equals the sum of the noninverting inputs minus those on the inverting inputs. Thus, we have a subtracting circuit. By making the two resistors in the feedback circuit larger, greater-than-unity gain may be obtained.

If we change resistor values, a **weighted** adder results, as shown in **fig. 17.** the feedback

fig. 15. Typical adder. Offset compensation is similar for all computing circuits.

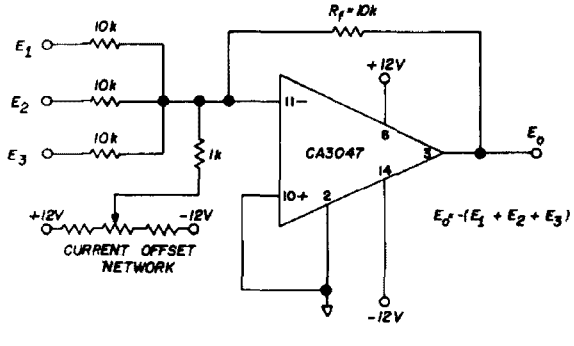


fig. 16. Adder-subtractor. Larger values of feedback resistors will result in a gain greater than unity.

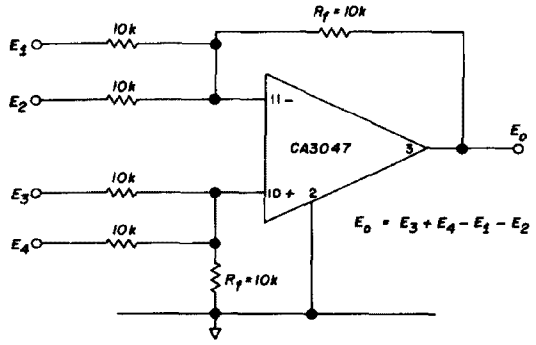
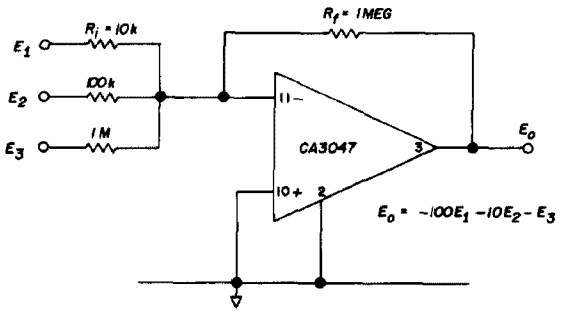


fig. 17. Weighted adder. The sum of the inputs is a function of the feedback resistor; any reasonable combination of Rf and Ri is permissible.



resistor value affects the sum of the inputs. The weight of the adder is proportional to input gain, which is determined by the feedback resistor.

Other mathematical operations in computers are integration and differentiation. The former is used to find the area under a curve; the latter determines the slope of a curve at any point. In the integrator of **fig. 18A,** a chopper-stabilized amplifier should be used because of the offset error caused by the feedback capacitor. If not chopper stabilized, a feedback resistor must be used.

Gain response of the integrator is maximum at the low frequencies and decreases linearly with increasing frequency. Amateur application of such a circuit would be in a lowpass filter following a speech clipper to attenuate harmonics.

In the idealized differentiator, gain increases indefinitely with frequency. To eliminate high-frequency noise problems, gain limiting is provided by R_i in the circuit of **fig. 18B.** This circuit is also useful for filter applications; frequency response is determined by the RC constants according to the equations shown.

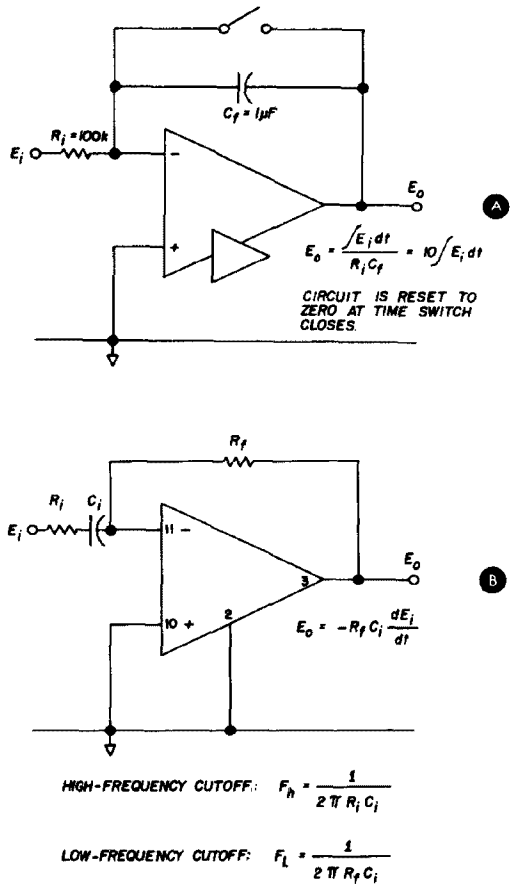
oscillators and waveform generators

Of the many operational amplifier circuits used in computers, probably the most popular amateur adaptations are oscillators and their close relatives, the multivibrators.

If you have a need for an oscillator with an unusually pure sine wave output, the Wien bridge[1] circuit in **fig. 19** is a good candidate. It is inherently temperature dependent, however. In the circuit shown, stability is improved with a lamp operating at very low current.

The multivibrator circuits in **fig. 20** have appeared in various forms in many amateur publications. They're used in electronic keyers, frequency counters, square-wave genera-

fig. 18. Integrator, A, and differentiator circuit, B. Double amplifier symbol in A denotes chopper stabilization required because of offset error due to Cf.



tors, and a host of other circuits where a controlled signal source is required.

The circuit of **fig. 20A** is an astable, or free-running multivibrator. Its uses include a timing-pulse generator, or clock, in counters. Feedback to the positive input is called "bootstrapping." This effectively increases circuit gain until it approaches infinity.

The bistable multivibrator (**fig. 20B**) has two stable states, each of which changes only when triggered by a pulse of opposite polarity. This circuit is used as a memory storage, counter, or shift register in computers. Its principles are often used in amateur circuits with little or no modification.

The monostable multivibrator, **fig. 20C,** is also called a one-shot. It has one stable state, which can be changed by an external pulse. It will then return to its original state after a time period determined by its RC constants. The one-shot is used for a time delay or to produce a pulse of specific width when triggered.

An application where the integrator feedback capacitor is allowed to charge is shown in **fig. 21.**[2] During a finite period, the input pulses will add algebraically until the amplifier saturates. When the switch is closed, the

fig. 20. Examples of the multivibrator. Circuit at A is free-running, or astable. B is a bistable multivibrator, or flip-flop. Monostable, or one-shot, is shown at C.

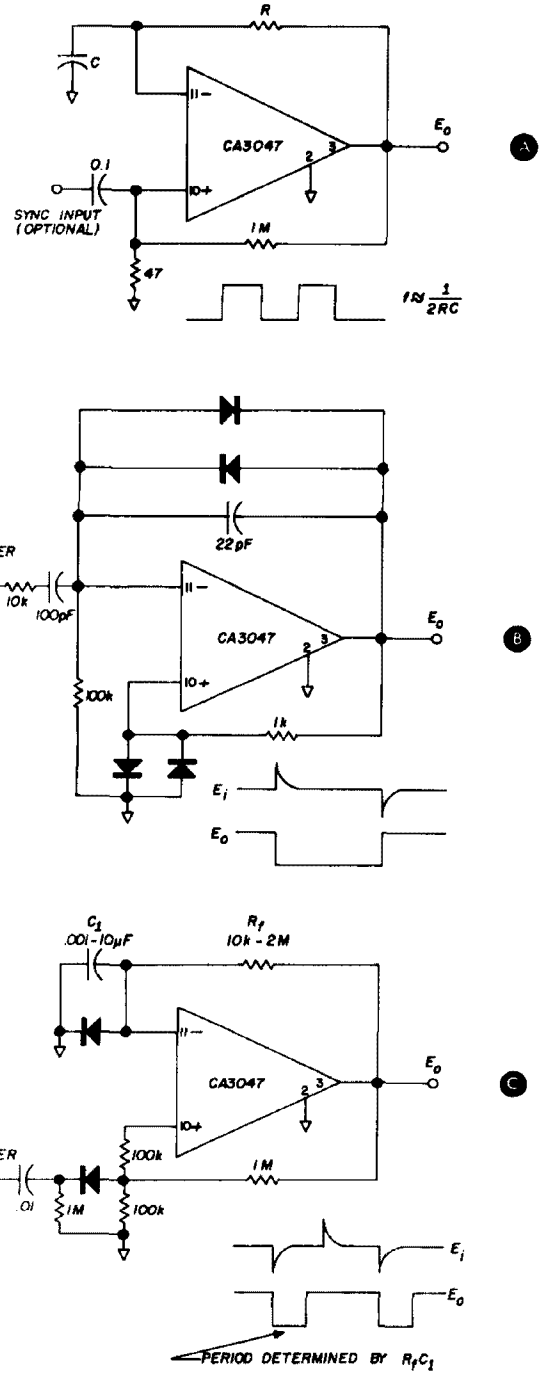


fig. 19. The Wien bridge oscillator. The circuit is temperature sensitive; stability is improved by the lamp, which operates at low-current level.

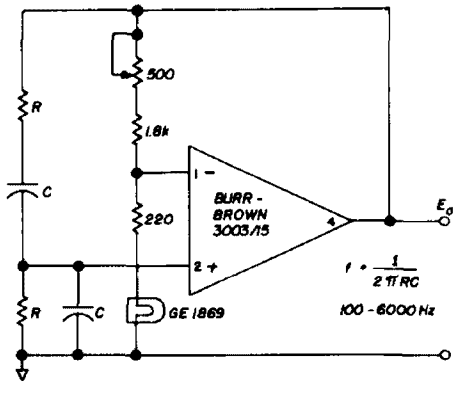
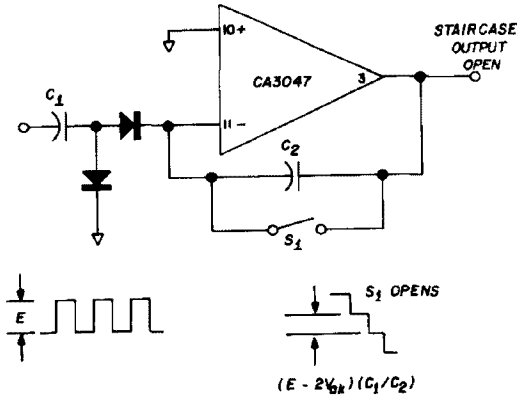


fig. 21. Staircase generator. Ramp output results if a dc signal is applied to pin 11 through a resistor.



output returns to zero. The circuit shown generates a "staircase" wave; it can be used as a ramp generator if a dc signal is applied to pin 11 through a resistor. Successively opening and closing the switch would give a sawtooth output.

Systems frequently require phase compensation for stability. Precise adjustment may be made with the technique shown in **fig. 22.** Adjustable lag is obtained by changing the input bypass capacitor; lead adjustment is provided by varying the feedback resistor. Resistors R_1 and R_2 may be necessary to stabilize the system.

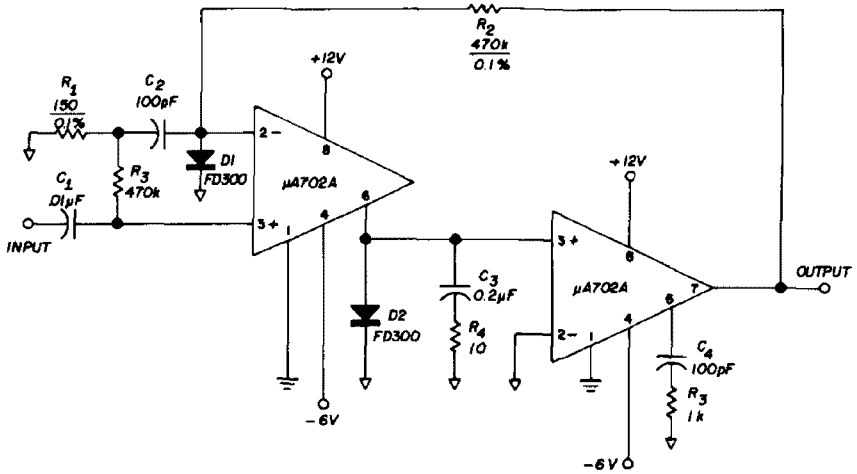


fig. 24. Precision ac amplifier. Gain is 70 dB; input impedance is 100 megohms.

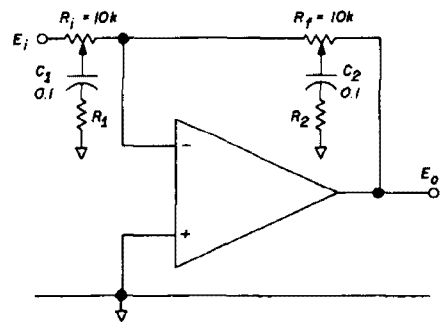


fig. 22. Phase-shift network for system stabilization. Lead and lag compensation are shown.

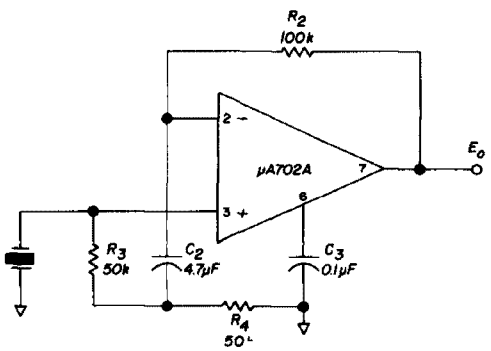


fig. 23. High input impedance amplifier for crystal transducer.

amplifiers

In addition to the basic amplifier circuits previously shown, I've included some useful variations.

The circuit of **fig. 23** is often used in dynamic instrumentation such as vibration measurements. It's a high input impedance amplifier using a crystal as a transducer.[3] A possible adaptation for amateur use would be a crystal microphone preamp.

The amplifier in **fig. 24** has a gain of 70 dB and an input impedance of 100 megohms.[3] Diodes are used to prevent latch-up. Because of the high-frequency characteristics (100 kHz) with the compensation shown, special attention should be given to layout and power-supply decoupling.

The tape-head amplifier[3] of **fig. 25** uses a matched pair of 2N3726's to reduce noise and increase input impedance. Despite the fact that it uses no input resistor (purists may object to classifying this circuit with op amps), the circuit does suggest a technique for improving common-mode rejection and increasing the common-mode range for any op amp.

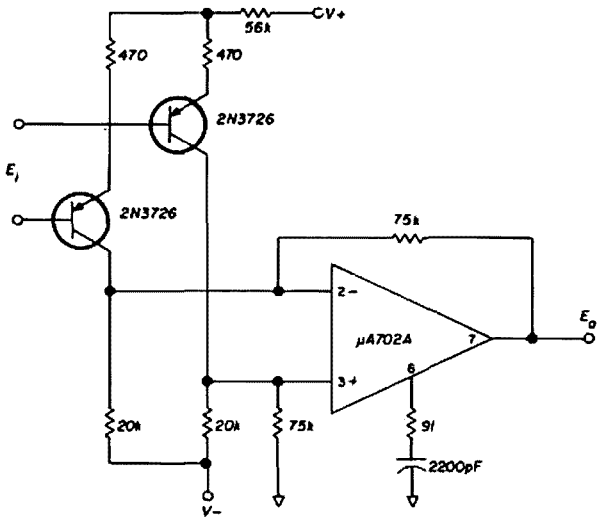
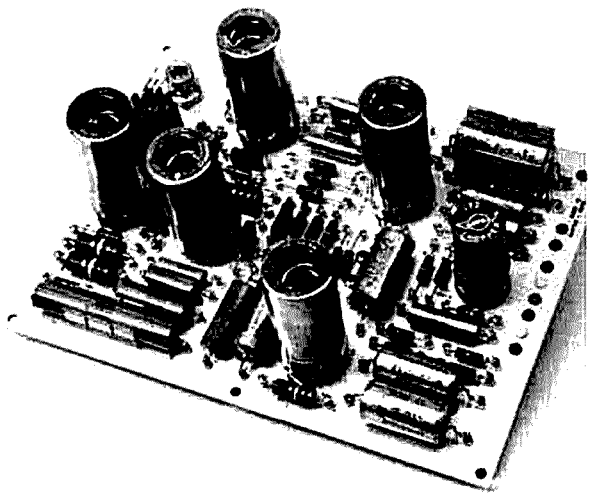


fig. 25. Low-noise tape head amplifier. Matchod transistor pair reduces noise and increases input impedance.



The Philbrick-Nexus USA4JT—"grandpappy of op amps." Very few amplifiers can match its performance. However, this fine unit has been retired to the back shelf because of its large size, aging characteristics, and high power consumption.

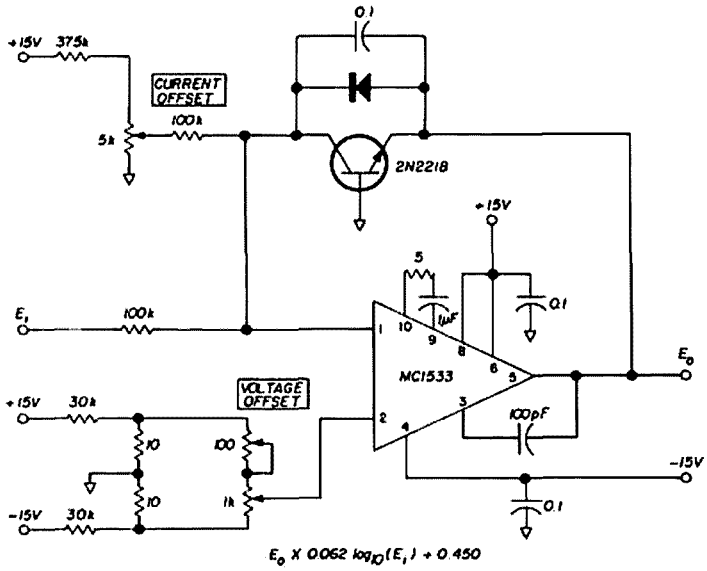
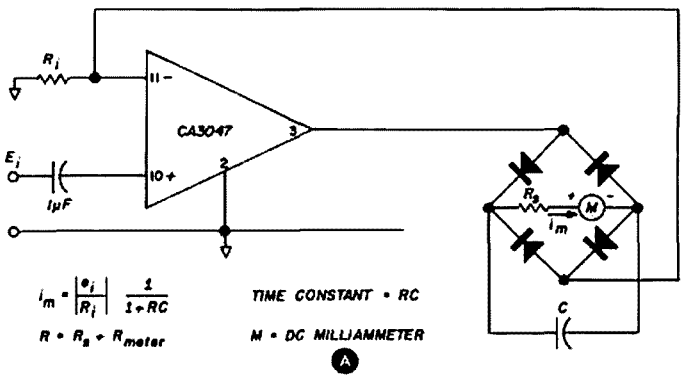


fig. 26. Logarithmic amplifier. Circuit operates over a frequency range of 6 decades.

measurement and control circuits

A widely used instrument is the log amplifier (or log converter). It has the capability of compressing input voltage ranges of several decades into a useful linear range. Some uses for this circuit (fig. 26) are in filter measurements, leakage measurements, and as a computer power-function generator. The amplifier shown uses a diode-transistor combination in the feedback circuit to achieve the conversion function.[1] Both current and voltage offsets are required for operation over a 6-decade input range. With an input of 0.13 mV to 100 V, the output is from 220 to 580 mV.

fig. 27. Lossless ac meter circuits. A high-impedance dc meter is preferred for the millivoltmeter circuit, A; a low-impedance meter should be used in the milliammeter circuit, B.



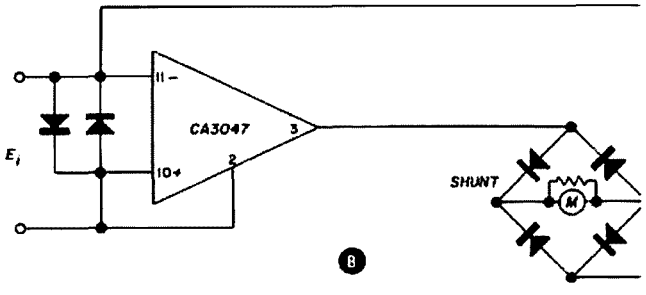


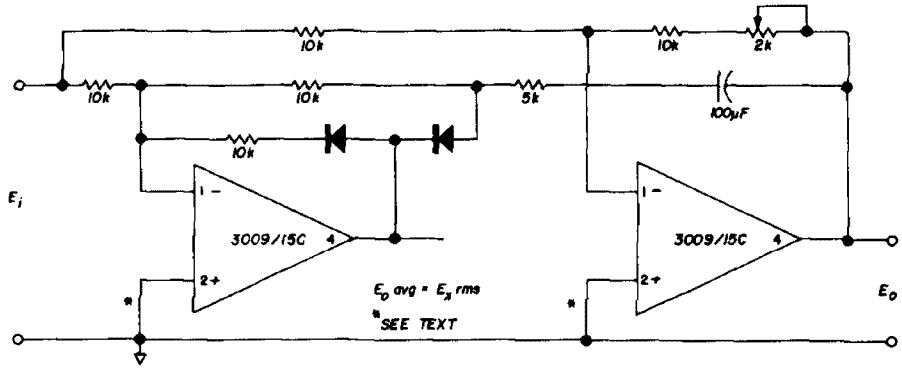
Fig. 27 shows two lossless ac meter circuits. The millivoltmeter circuit, **A,** uses an op amp to compensate for diode, resistor, and meter losses. The response time, which is usually low, can be increased by increasing either the meter series resistor, R_s, or the averaging capacitor, C.

The current-sensitive counterpart of the millivoltmeter, shown in **fig. 27B,** has zero drop across its terminals. Limiting diodes at the input should have very low leakage. No charging capacitor is necessary, because the current is averaged by the meter. Low-impedance dc meters are practical in this circuit, whereas the millivoltmeter performs more efficiently with a high-impedance meter.

In measurement and control circuits, it's frequently necessary to convert ac to dc. The circuit of **fig. 28** using Burr-Brown amplifiers[1] consists of a full-wave rectifier and a filter.*

The time-delay circuit in **fig. 29** requires an amplifier with a high-impedance input such as that provided by an fet. The Burr-Brown 1552/15 is such an amplifier. Of the many uses for this circuit in amateur applications, an example would be to control timing of voltage turn-on in a power supply. Circuit response time would be limited by relay action.

fig. 28. Ac-to-dc converter. Circuit has an input range from 6 mV to 6 V rms at 10 to 1000 Hz. Amplifier is a Burr-Brown IC.



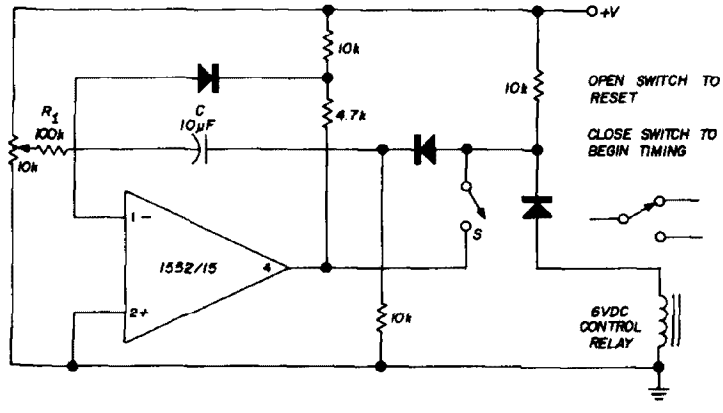


fig. 29. Timing control circuit. Minimum delay is determined by R1-C; maximum delay is infinity.

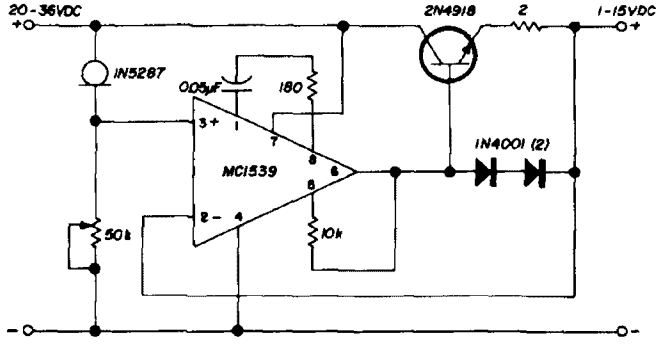


fig. 30. A power supply that provides up to 200 mA between 1 and 15 volts. Regulation is better than 0.01%!

power supplies

The Motorola MC1539 op amp is the center of precision in the circuit of **fig. 30.**[5] This supply provides up to 200 mA at any voltage between 1 and 15 volts. Regulation is better than 0.01% The unusual reference supply consists of a constant-current diode (type 1N5287) and a 50 kilohm potentiometer.

The amplifier has a gain of 120,000, so it won't load the reference. Note that output compensation is between pins 5 and 6, and input compensation is between pins 1 and 8. The circuit is protected against burnout from short circuits.

*Although not shown in the schematic, each amplifier should have a resistor between pin 2 and ground. These must be chosen to minimize output errors due to input offset current. The resistor for the second amplifier will be critical because of changes in circuit impedance as a function of input voltage swing. ***Editor***

voltage reference

A more sophisticated reference supply uses its own op amp, a National Semiconductor LM101 (**fig. 31**).[6] The 1N827 reference diode is temperature compensated. Regulation is 0.01 mV/V, and temperature stability is ±0.05% from −55°C to 125°C. Short-circuit protection for the reference is provided internally. The LM101 needs only one compensating component, the 33-pF capacitor.

reversible, precision power supply

This supply, **fig. 32,** is a classic representation, because many basic concepts can be demonstrated by analyzing its characteristics.

A Fairchild A00-7 amplifier, referenced to a stable source, drives a dual buffer that increases the output-current capability. By reversing the reference-supply polarity, the output reverses and switches to the opposite half of the buffer.

Digital programming of voltage is practical by changing the values of R_f and R_i. The system shown is limited to a gain of unity, which means that the output is never less than the reference voltage.

Several important parts of the circuit should be noted. First, the op amp must have a larger voltage swing than the buffer output voltage. The op amp and buffer are compensated separately. This may not be possible if there's much wire between them.

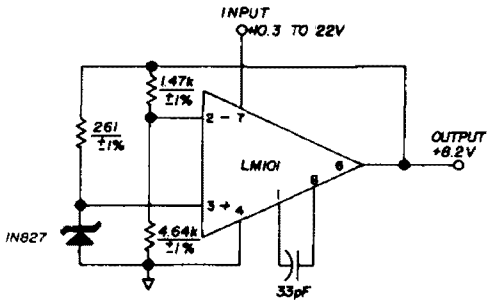
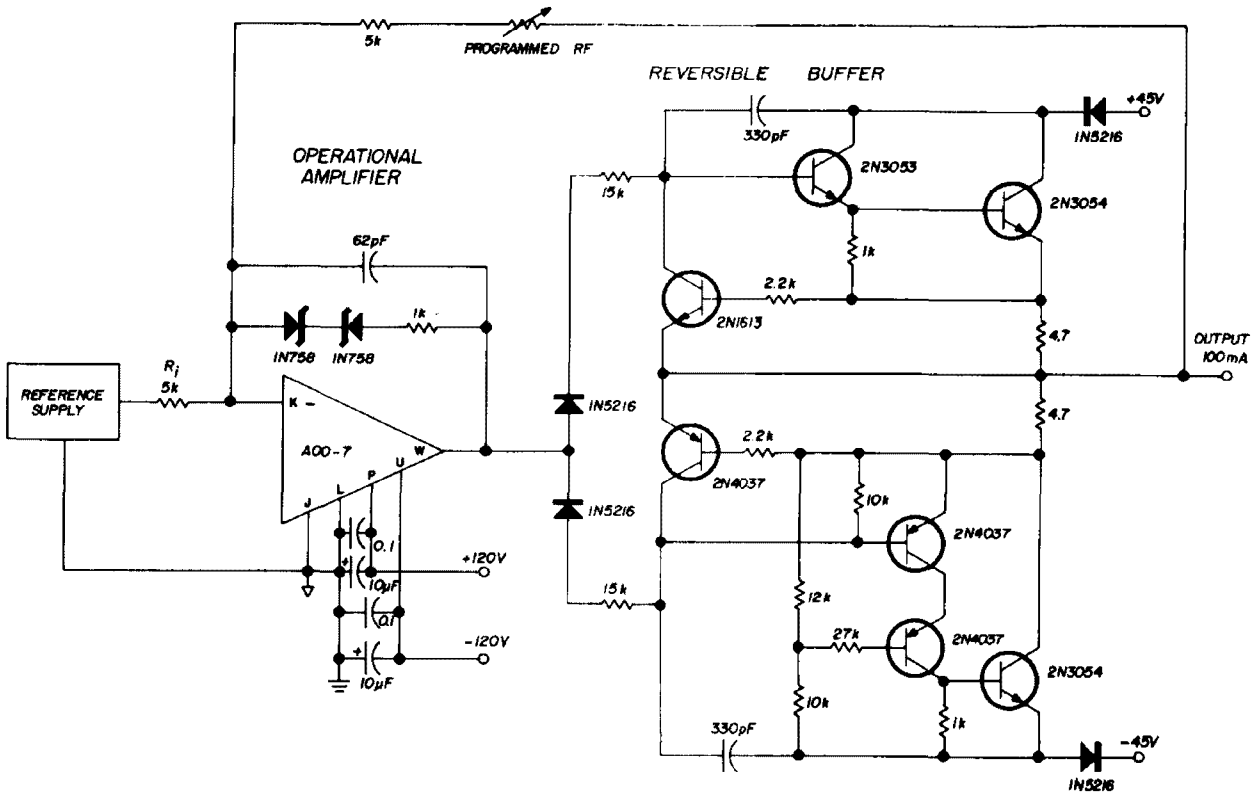


fig. 31. Voltage reference source. Regulation is 0.01 mv/V; temperature stability is ±0.5% from −55° to 125° C.

Another interesting feature is the zener diode clamp in the op amp feedback circuit. This prevents the amplifier from locking in saturation. Back-to-back zeners limit input voltage swing in either polarity. The

fig. 32. Reversible, precision power supply. Digital programming of voltage is practical by varying Rf and Ri; zener diode clamp in feedback circuit prevents locking in saturation.



buffer is protected against short circuits by a transistor limiter. This cuts off the pass transistors when the current through the 4.7-ohm resistor becomes excessive. Output of the A00-7 is limited by the 15k resistor.

active filters

A nice thing about active filters is that you don't need inductors to achieve near-ideal mathematical response characteristics. Another good feature is high input impedance, which means that matching is not a consideration.* While compensation for the operational amplifier is necessary, filter reactance trimming is not. Once you've calculated component values for a specific response, you're done.

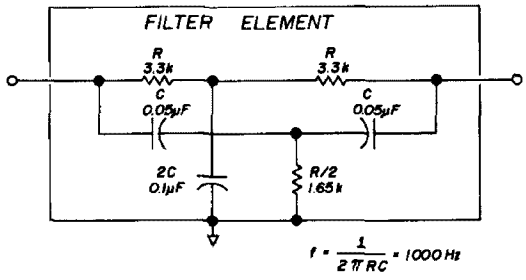


fig. 34. A Twin T filter for use in an op amp circuit. Bandpass is 1000 Hz; with input and feedback resistors of 10k and 100 the gain would be 10. The product CiRi should be greater than twice CR.

A circuit that's easy to understand is shown in **fig. 33.** A filter in the feedback circuit of a conventional inverting amplifier yields an output with the response characteristic of the filter.

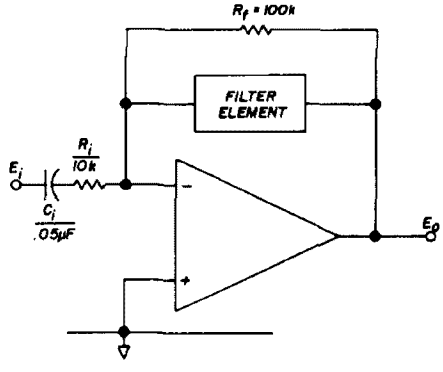


fig. 33. The op amp in a filter circuit. Filter in the feedback circuit yields an output with response characteristic of the filter element.

A possible filter is the Twin T shown in **fig. 34.** A 1000 Hz bandpass filter is in the basic circuit. If $R_i = 10k$ and $R_f = 100k$, the gain will be 10 at 1000 kHz. The Twin T is one of the simplest (first-order) filter elements; however, it has relatively low Q, so don't expect miracles from it.

The circuit of **fig. 33** may be used with an active high pass, low pass, or rejection-notch filter by inserting the appropriate filter element. Reference 7 provides more information for active filter designs.

Another practical approach toward building an improved filter is to precede a conventional filter with an op amp follower (**fig. 35**). This circuit eliminates filter input loading problems. Although resistor R is chosen to

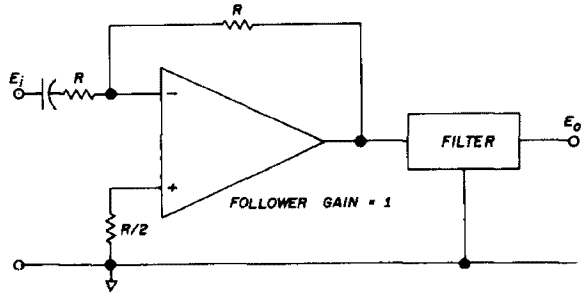


fig. 35. Preceding a conventional filter with an op amp follower to eliminate input loading. Input impedance of amplifier is arbitrary, as explained in the text. Resistor R should equal the input impedance of the filter.

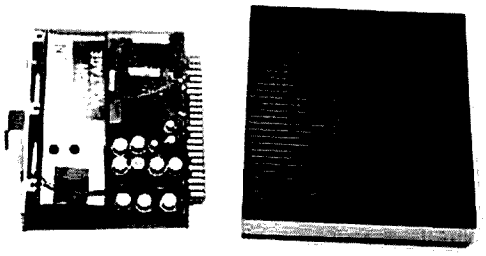
equal the filter input impedance, the resistor is really used to match the input of the preceding stage. The input impedance of the op amp is arbitrary.

A follower on the filter output would be useful if a varying load is used. The purist will argue that this isn't a true active filter. I'm willing to concede the point, but I hasten to add that it's a **handy technique.** I encourage the amateur to take it from here.

a parting thought

I've presented some basic data on one of the most interesting and challenging products of modern solid-state technology. The circuits shown are the most commonly used, but by no means do they cover the entire

*True for this circuit, but not for controlled source and negative immitance converter techniques. **Editor**



A noteworthy successor to the best tube op amps is the Fairchild A00-7. It features discrete components, built-in compensation, and chopper stabilization.

field of possible applications.

If you wish to adapt these circuits to your needs, a good grasp of op amp theory is essential; the material listed in the references will supplement that in the first part of this article. Some possible projects that come to mind are:

1. An ultra-stable oscillator (for system synthesis).
2. A precision filter for selective calling.
3. A high-impedance meter for measurement of h_{fe} or g_{fs}.
4. A precision digital power supply.
5. Science Fair computer projects.

I'm sure you've thought of a few projects, too.

references

1. "Handbook of Operational Amplifier Applications," Burr-Brown Research Corp., Tucson, Arizona.
2. H. A. Wittlinger, "Application of RCA CA3033 and CA3033A, etc.," RCA, Electronic Components, Harrison, New Jersey.
3. "Linear Applications Manual," Fairchild Semiconductor, Mountain View, California.
4. "Integrated Circuit Data," Motorola Semiconductor Products, Inc., Phoenix, Arizona.
5. "Motorola Monitor," Vol. 7 No. 1, Motorola Semiconductor Products, Inc., Phoenix, Arizona.
6. "Ideas for Design," *Electronic Design*, January 18, 1968, p. 128.
7. "Handbook of Operational Amplifier Active RC Networks," Burr-Brown Research Corp., Tucson, Arizona.
8. "Application Manual for Computing Amplifiers," Philbrick-Nexus, Dedham, Massachusetts.

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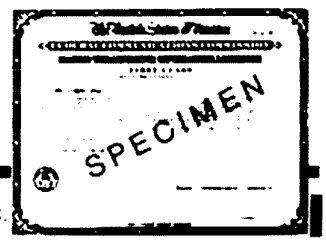
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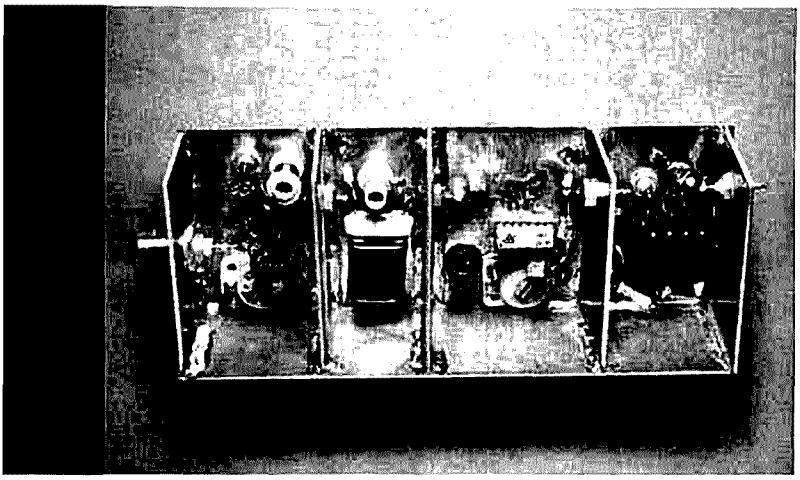
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a fixed-tuned receiver for wwv

Hank Olson, W6GXN, P. O. Box 339, Menlo Park, California 94015

Many modern amateur receivers are ham-band-only types and must have an adapter to receive WWV. Even if you're using a general-coverage hf receiver, it is quite a nuisance to go "way up the band" to get WWV so you can set your 100-kHz crystal calibrator. Wouldn't it be convenient to have an entirely separate, fix-tuned WWV receiver **only** to check the calibrator? The simple receiver described below fills the bill with just a few components.

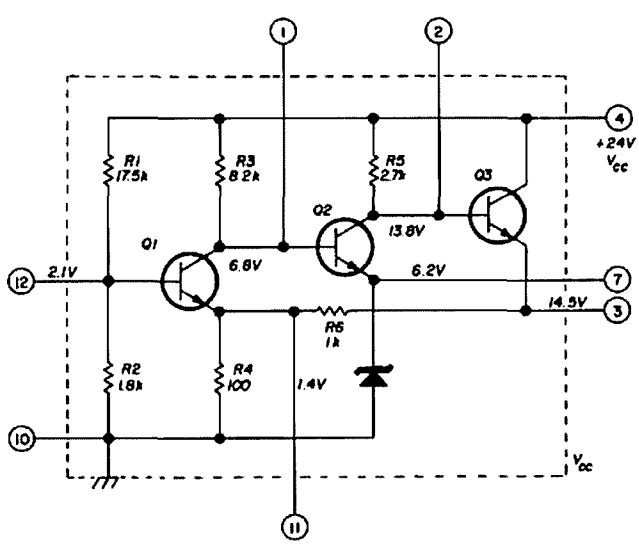
This WWV receiver is designed for 15 MHz, but it could be used on 20, 10 or even 5 MHz with changes in the resonant circuit elements. (For receiving WWV on 2.5 MHz, it is probably easier to modify a transistorized broadcast receiver.)

The majority of U. S. amateurs probably use the 10- or 15-MHz transmissions of WWV or WWVH, especially during daylight hours; so 15 MHz was chosen for this receiver.

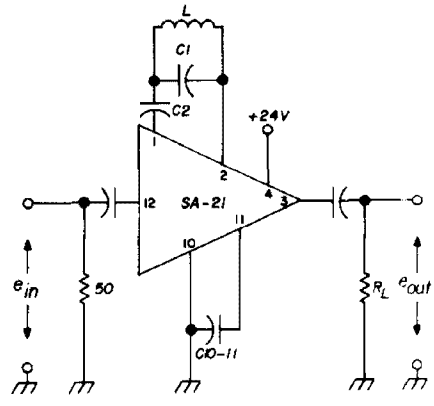
the SA-21 integrated circuit

The heart of the WWV receiver is a Sylvania SA-21 video amplifier. This integrated circuit has a useful frequency range that extends beyond 100 MHz, which is quite a bit more bandwidth than many of us are accustomed to calling "video." Used as a narrowband amplifier, the SA-21 provides over 50 dB gain at 15 MHz, as shown in **fig. 2.**

fig. 1. Sylvania SA-21 internal circuit. Voltages are nominal quiescent values at 25° C.



Unlike most linear microcircuits, the SA-21 is **not** a differential transistor amplifier; its internal circuit is shown in **fig. 1.** Both figs. **1** and **2** were taken from reference 1.



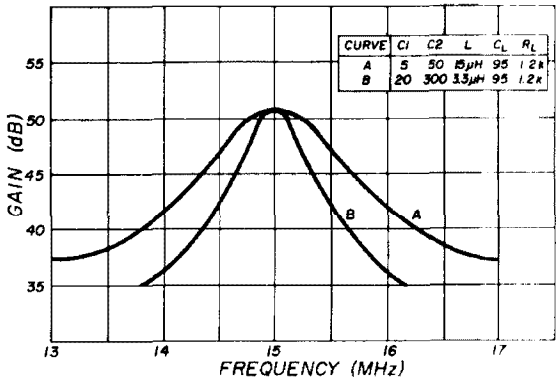


fig. 2. Bandpass response of the Sylvania SA-21 wideband amplifier. Curve B was used in this design.

the receiver circuit

The WWV receiver schematic is shown in **fig. 3.** An rf stage and simple crystal filter are used ahead of the IC amplifier to provide the selectivity required in such a trf circuit. An fet is used as an rf amplifier, because it has inherently good cross-modulation immunity and provides a higher input impedance in the common-source configuration than a bipolar transistor in the common-emitter configuration. The higher input impedance of the fet means that the step-up ratio of the antenna transformer, L1, must be large, which in turn produces higher operating Q (better selectivity).

crystal filter

The crystal filter is very simple. A slug-tuned inductance in parallel tunes out holder capacitance. The resulting filter circuit is then as shown in **fig. 4.**

A 15-MHz crystal which is cut for series resonance should be used. I tried an old CR18/U crystal, cut for 32 pF operation, with partial success. One would expect this crystal to have its passband on the low side of 15 MHz, but the passband center frequency was at 14.998 MHz (2 kHz low). This was close enough to allow intelligible copy of WWV, since the filter bandwidth is about 6 kHz. Replacement of the CR18/U crystal provided somewhat better fidelity.

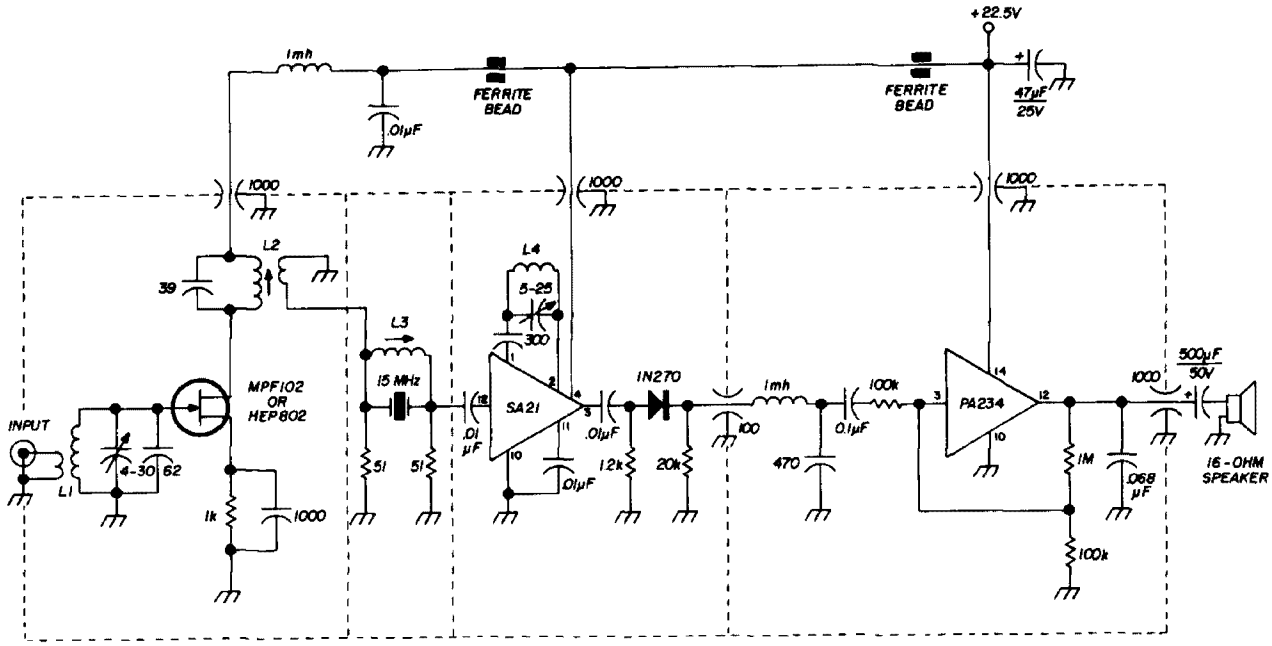
detector and audio stages

Following the SA-21 amplifier is a simple diode detector, a 1N270, and an IC audio amplifier, PA234. The General Electric PA234 is the least expensive of a family of their amplifiers IC's, PA222, PA234, PA237 and PA246. The PA234 will provide up to 1 watt of audio. It is a modified dual in-line package (DIP). Every other pin is omitted on the PA234. Also, a heat sink tab protrudes from the end containing pins 7 and 8. This makes for easy PC board mounting (lots of room between pins). However, in this particular receiver, both IC's are mounted in 14-pin DIP sockets.

construction

The receiver is partitioned into four shielded sections: rf amplifier, crystal filter, SA-21 and detector, and audio amplifier. Power for each section (except the crystal filter) is brought out through the top plate via 1000-pF feedthrough capacitors. Decoupling the +22.5-volt supply bus is quite important (don't try any shortcuts)! Decoupling elements are ferrite beads and an rf choke. The Ferroxcube ferrite beads (K5-001-00/3B) present about 50 + j50 ohms at high frequencies; so if you **must** substitute, bear this in mind.[2]

The IC's were mounted in plastic DIP sockets made by Methode (M1141). These are particularly suitable, because each has



L1 1.35 μH, 19 turns no. 20 on Micrometals* T50-10 core, with 3 turns no. 22 primary

L2 2.0 to 3.7 μH, CTC X2060-1 with 4 turns no. 26 secondary

L3 10 to 18 μH, C.T.C. X2060-4

L4 3.3 μH, 30 turns no. 26 on Micrometals* T44-10 core

fig. 3. WWV fixed-tuned receiver schematic. Note shielded compartments and decoupling circuit.

two mounting holes between pin rows. These holes can be tapped for 4-40 screws and the sockets mounted in a stand-off configuration with two small ceramic spacers (left over from the last Centralab wafer switch you assembled). This puts the IC socket pins down near the ground plane for low-inductance grounding or bypassing. The socket for the PA234 is crowded next to the end plate so the heat sink tab may be easily soldered.

The entire assembly is made of sheared epoxy-glass, double-sided, copper laminate of the type used to make PC boards. This material is easily soldered into a rigid rf package (no solder lugs are needed, saving space). The rf conductivity is superior to that of an aluminum chassis. I found that the bottom plate and one side could be left off without causing instability, so the chassis work is complete as shown in the photo.

the toroid coils

A word is in order about L1 and L4. They are wound on powdered iron toroids. These particular toroids are **made for the frequency** and give substantially better performance than most other coils (except perhaps miniductor or Airdux types used in open space). The curves of Q vs frequency for the two toroids are shown in **fig. 5.**

Another advantage of these powdered iron toroids (made by Micro Metals)* is their lack of temperature sensitivity. Unlike ferrite cores, these toroids have remarkable stability of both inductance **and Q** with temperature. If you decide to use coils that are not toroids to replace L1 and L4, more shielding may be required to prevent oscillation.

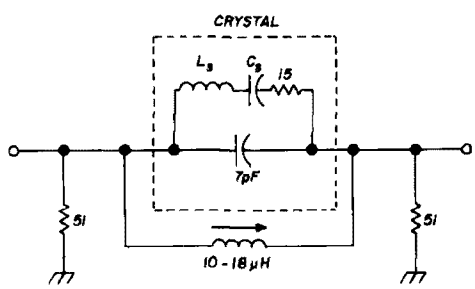
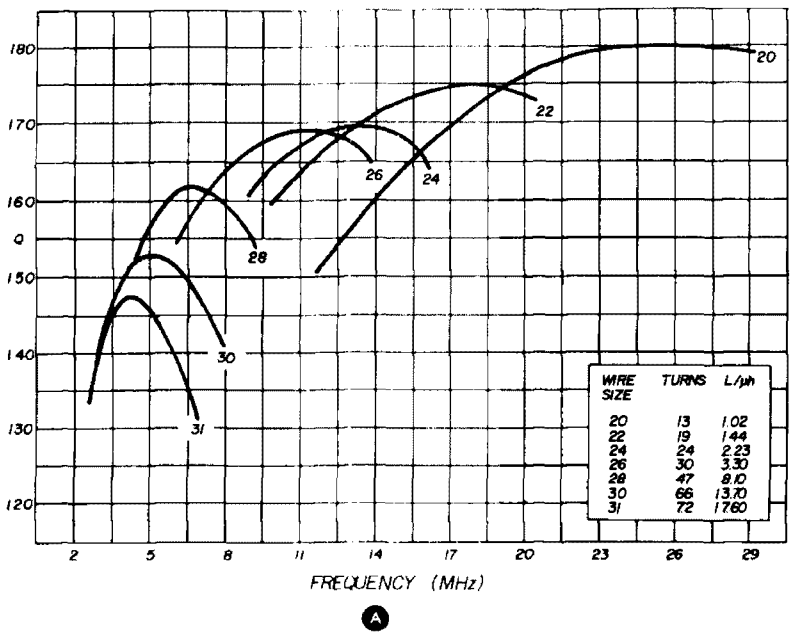


fig. 4. Equivalent circuit of 15-MHz crystal filter. Crystal is AT cut, series resonant.

*Micrometals, 72 E. Montecito Avenue, Sierra Madre, California 91424



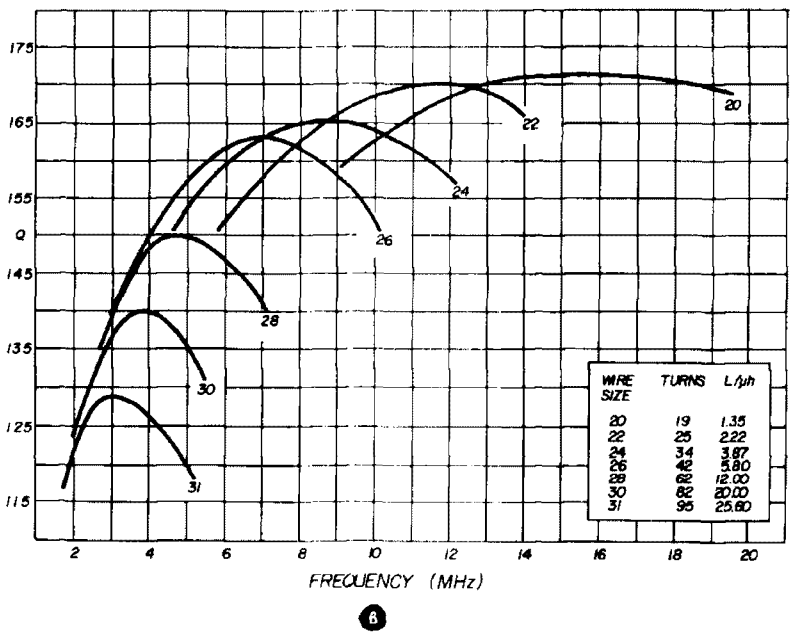


fig. 5. Operating Q as a function of frequency for the Micro Metals T44-10 toroid, A, and the T50-10, B. L1 and L4 were constructed from this data.

operation

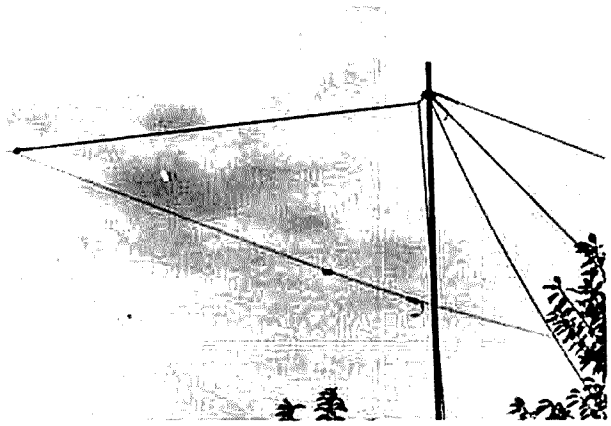
The unit is powered by a 22.5-volt battery (Burgess 4156) and idles at about 25 mA. Strong signals will push current consumption up as high as 75 mA, but that's getting pretty loud!

No agc or gain control is included in this unit; a variable attenuator is used ahead of the rf amplifier.[3] During the day at my location, about 20 dB of attenuation is used. You might argue that this is a crude approach; but let's face it, that's the best place to control gain in a fixed-tuned receiver.

references

1. "Wide Band Video Amplifier, SA-20 Series," Sylvania Data and Application Sheet, February 15, 1967.
2. H. Olson, "Ferrites, or What's Mu with You," CQ, April 1966, pp. 49-52.
3. G. Daughters and W. Alexander, "Low-Power Attenuators for the Amateur Bands," *73*, January 1967, p. 40.

ham radio



a multiband long-wire antenna

Design details for a low-cost efficient radiator

Edward M. Noll, W3FQJ, 3510 Limekiln Pike, Chalfont, Pennsylvania 18914

Would you like a multiband antenna that favors the direction of hard-to-get states or countries, yet has good omnidirectional coverage? One answer is the long wire, a much neglected antenna that deserves more recognition.

The long wire is mentioned often in the literature. Information on feed methods is meager, however, which may be why it's seldom used. If you have about 300 feet of reasonably clear space, you can erect the long wire described here and enjoy excellent performance on 40 through 10 meters. Depending on the feed method, the antenna gives either unidirectional or bidirectional coverage, with substantial gain in the direction of the wire. Power gain over a dipole is from 1.2 to 7, corresponding to lengths from 1 to 12 wavelengths. (Numbers are ratios, not dB.) Minor lobes, concentrated near the center of the wire, provide omnidirectional response.

The single long wire is probably the simplest antenna you can build that will provide maximum gain for lowest cost and effort. Only two masts are required; however, one will do (at the far end) if you use your eaves

to support the near end. The tilt thus provided increases radiation at the low vertical angles for DX, as discussed below.

principles

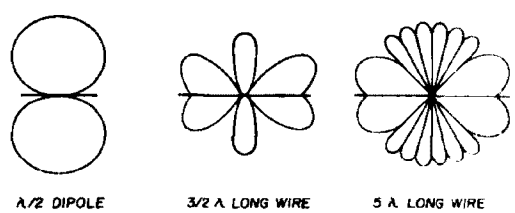
The dipole, which is one-half electrical wavelength long, is the simplest resonant antenna. Two major lobes occur at right angles to the antenna (**fig. 1**). If the wire length is increased in multiples of one-half wavelength, the antenna will also be resonant at these points. Additional lobes appear as the wire length is increased. The major lobes move in the direction of the wire, so that maximum radiation is concentrated off the end.

The longer the wire, in multiples of one-half wavelength, the greater will be its gain in the direction of the major lobes as compared to a dipole. The secondary lobes, **fig. 1**, give good omnidirectional coverage.

feed point

Resonant long-wire antennas can be center- or end-fed (**fig. 2**). A low-impedance point is made available at the center by making each leg an odd multiple of a quarter-wave-length long. Contrary to a rather wide belief that the precise length of a long-wire antenna is unimportant, the leg lengths must be accurately cut to obtain a low-impedance feed point.

fig. 1. Horizontal radiation pattern of long wire as a function of length.



A low-impedance feed point for end feed is found a quarter wavelength in from the transmitter end of the long wire. End feeding is a preferred method for two reasons. The antenna is more unidirectional with a higher gain in the direction of the long leg. The longer the long leg, the higher the gain and the more directive the pattern becomes off its end.

It is apparent from the above that a long-wire antenna can be erected to favor a given area. If a group of states give you trouble, point the antenna in their direction.

If you prefer a bidirectional pattern, use a center feed. If you live along the East Coast, the long leg can be pointed west. Conversely, for a western station, the long leg can be directed east. In the Central States you may prefer a bidirectional center-fed type, or perhaps you might want the end-fed unidirectional characteristic if you are having difficulties with certain states.

antenna length

Each leg must be an odd multiple of a quarter-wavelength for center feed.

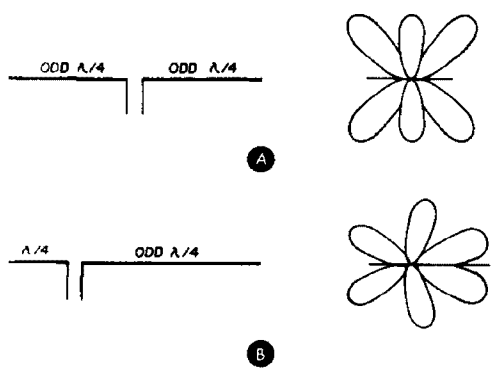


fig. 2. Center feeding the long wire, A, and end feed, B, with respective patterns.

$$\text{leg length} = \frac{246\ (N)}{f\ \text{MHz}}$$

Where N is the number of quarter wavelengths.

End effect is important on these long wires, and you must trim the legs evenly to reach a given resonant frequency. Start with the formula given above, and prune until resonance is obtained. Usually, the antenna must be shortened from two to six percent as a function of antenna height, tilt, and nearby conducting surfaces.

Antenna resistance is low and varies slowly with length, **fig. 3.**[1] **Fig. 4** shows two end-feed methods for different antenna lengths. Using the coax cable impedance shown, you can obtain a match with an swr of about 1.5:1 without a tuner.

The length of the short leg of the end-fed type can be calculated using the regular dipole equation. The long leg must be made some multiple of a quarter wavelength.

$$\text{long leg} = \frac{246\ (N)}{f\ \text{MHz}}\ (N)\ \text{feet}$$

$$\text{short leg} = \frac{234}{f\ \text{MHz}}\ \text{feet}$$

Again, the antenna legs must be trimmed carefully to find resonance and establish a feed-point impedance that can match the transmission line. Short sections should be trimmed off the quarter-wave segment to obtain resonance just as you trim an ordinary dipole. Because the long leg is so very long, you can trim off larger pieces of the antenna wire in moving toward the desired resonant point.

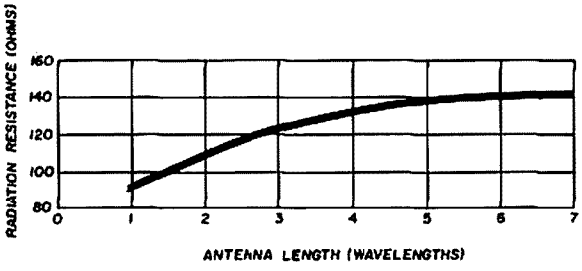


fig. 3. Change of radiation resistance with length of longwire antenna.

For the end-fed long-wire, it is helpful to first set the quarter-wave dipole leg to resonance. Then the longer leg is trimmed for a minimum swr. Two useful instruments for trimming this antenna are an swr meter and an antenna noise bridge. Always connect an swr meter in the line at some point that is a whole multiple of an electrical half wavelength away from the antenna feed point.

In fact, it is advisable to use a transmission line that is a multiple of a half wavelength between antenna and transmitter. In so doing, the low resistive impedance of the antenna is reflected to the transmitter, and the transmitter need not cope with reactive components that can be introduced by the line, however small its standing wave ratio.

multiband operation

For optimum results the long-wire antenna should be resonated on each band. There are three ways to do this. A tuner can be mounted at the feed point, but this presents a weather-proofing problem.

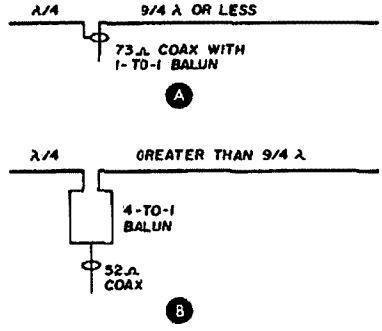


fig. 4. Two methods of feeding a long wire at the end. A swr of about 1.5:1 can be obtained using three systems.

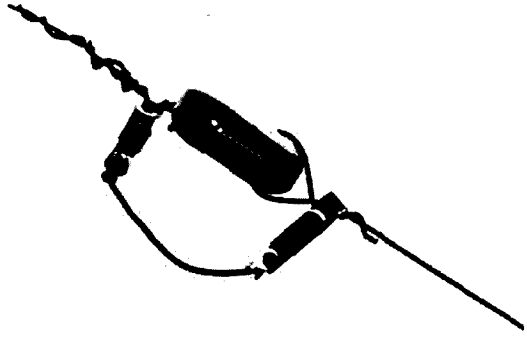
Furthermore, the tuner must be retuned when changing bands or when shifting operation from one part of the band to another.

Reasonable results can be obtained by using a tuner at the transmitter. However, this technique makes the transmission line a part of the antenna, and the antenna pattern can be affected adversely. Furthermore, a high swr can develop on the line. Recall that a tuner at the transmitter actually works for the transmitter and does very little for the antenna system.

The third possibility is to preset the long-wire antenna for each band. This can be done by bringing the antenna ends down near the ground level (**fig. 5A**). This causes no serious pattern change. The antenna can be trimmed carefully for each band to obtain optimum operation without the use of any tuner.

For DXing the long-wire can be tilted slightly in the direction of the long leg (**fig.**

Insulator-jumper combination.

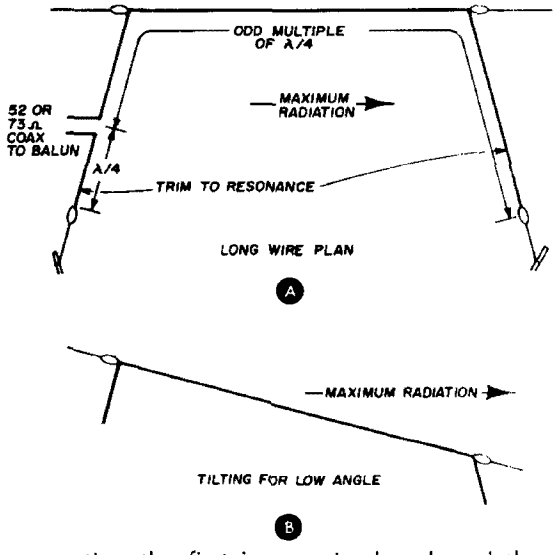


5B) to improve the low-angle radiation in the favored direction. The angle increases in the opposite direction.

W3FQJ end-fed long wire

The arrangement and dimensions of a practical antenna are given in **fig. 6.** The quarter-wave dipole segments are easy to set up and permit 10- through 40-meter operation. A band change is made quickly by proper arrangement of the jumpers at each end of the antenna. For 20-meter operation the first jumper is left open. For 15-meter

fig. 5. To avoid use of tuner, long wire ends are brought near ground and carefully trimmed for resonance, A. Antenna bandwidth will be quite narrow. At B, tilting the wire improves low-angle radiation slightly in direction of tilt.



operation the first jumper is closed, and the second jumper is opened. On 20 meters the first two jumpers are closed, and the third jumper is opened. All jumpers are connected for 40-meter operation.

The long leg of the antenna is 9 quarter-wavelengths long on 40, 17/4 on 20, 25/4 on 15, and 33/4 on 10. Formula values are

$$(40)\ \text{leg length} = \frac{2214}{7.2} = 307 \text{ feet}$$

$$(20)\ \text{leg length} = \frac{4182}{14.2} = 294 \text{ feet}$$

$$(15)\ \text{leg length} = \frac{6150}{21.3} = 288 \text{ feet}$$

$$(10)\ \text{leg length} = \frac{8118}{28.6} = 283 \text{ feet}$$

After trimming, the practical lengths reduce to 297 feet 7 inches, 271 feet, and 272 feet 10 inches respectively. Note that the same leg length can be used for both 10 and 15 meter operation. Thus in changing operation between these bands, only the near-end jumpers need be shifted.

results

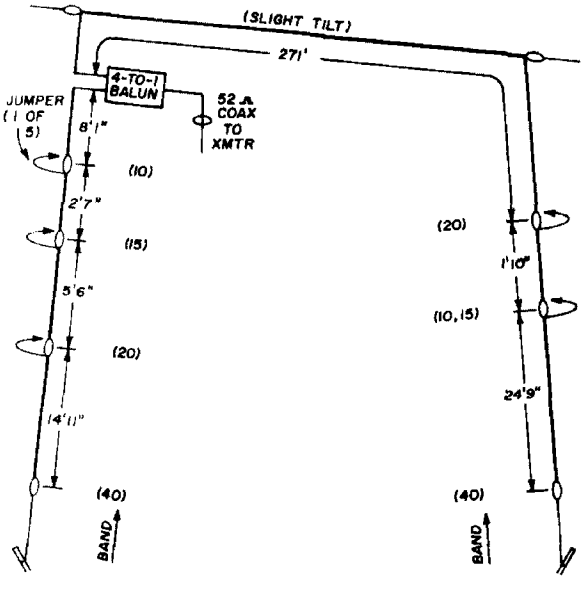
Results have been gratifying on all four bands. The bearing of the long leg, as erected here in Eastern Pennsylvania, is set at 255°. On 10 meters, where the directivity is sharpest, this places a strong-signal area diagonally across the Continental United States. At the same time, good reports are obtained in the Southern States and in North Central States, thanks to the secondary lobes. Of course, on the lower bands the number of electrical wavelengths on the legs is not as great, and the horizontal radiation pattern is less sharp, encompassing a larger area of major-lobe coverage.

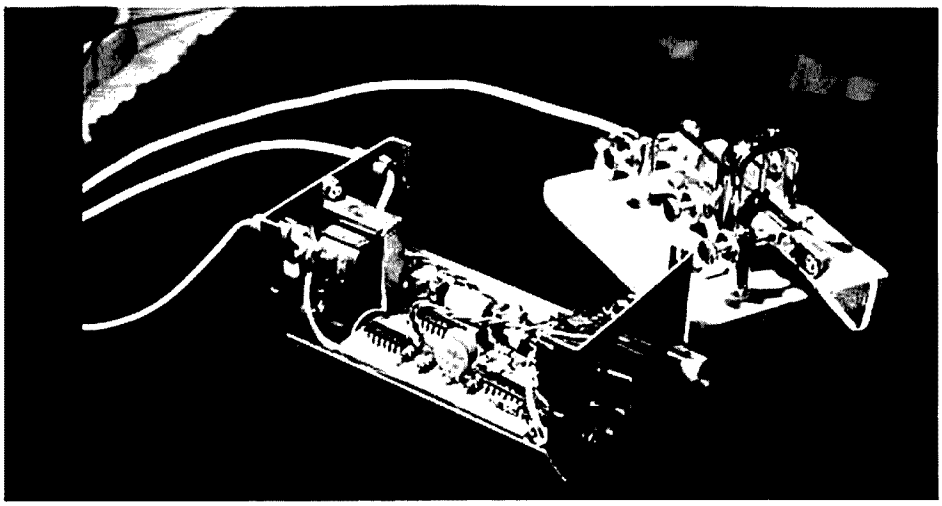
references

1. "Antennas and Radio Propagation," Dept. of the Army Technical Manual TM 11-666, February, 1953.

ham radio

fig. 6. The multiband, segmented long wire at W3FQJ. Coax cable should be at a right angle to feed point if possible.





one more electronic keyer

A dot memory ensures positive character formation in this improved version of the TO keyer

G. D. Young, VE7BFK, 14409 West 11th Avenue, Vancouver 8, British Columbia

During the past few years several articles have appeared describing the electronic key.[1,2,3] Here's one more. This key is a TO type and is basically the same as the one described by K3CUW in *QST*.[1] Its features are:

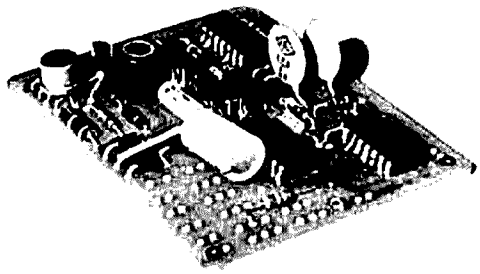
1. A dot memory
2. Small size
3. Npn or Pnp transistor switching
4. "Squeeze-key" operation if desired.

I'll start by offering some motivation for its design, followed by a review of the integrated circuit (IC) characteristics.

I first became familiar with electronic keys with the construction of an all-transistor TO key,[4] which I used extensively for about five years. In constructing a new key, I felt I should strive for increased ease of operation and a substantial reduction in size. (The key mentioned above occupied 3½x6x8 inches on the operating table.)

For easier operation, an electronic key must have a memory. Keys have been described with dot and dash memories,[5,6] but I decided that a key with a dot memory would only be best for the reasons given below.

If you now use an electronic key with no



Main circuit board.

memory, you know that one of the most common errors is the omission of a dot in letters such as n, k, c, y, etc. This error can be corrected if a dot memory is included, which will insert the dot if the paddle is closed during the preceding dash.

Because it's longer, a dash followed by a dot is seldom omitted. The natural tendency to hold the paddle on the dash side for a longer time ensures that the dash will be formed.

A dash memory would be required if the key had a free-running clock, as in the Ultimatic.[5,6] However, I felt that a **gated** clock would make the key easier to operate, because the response to touching the paddle is immediate. The free-running clock key may delay up to a whole dot duration before doing anything.

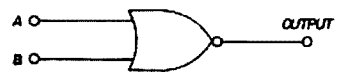
The small size requirement suggested, (1) integrated circuits and, (2) that something be done about the huge filter capacitors ordinarily required for low-voltage, high-current power supplies. An electronically regulated supply takes care of the latter problem.

review of ic operation

The IC's used in this keyer should be in fairly common use by now.[7] For completeness, however, I'll include a brief description of their operation. Rather than consider IC operation in terms of Boolean functions, I'll regard them as black boxes with particular characteristics. I believe it's less confusing to use this approach for a practical understanding of their operation.

The symbols and names are borrowed from the language of Boolean algebra (for example, the gate is a **NOR** gate). But it's best to understand the way in which the circuit works rather than the logic functions involved.

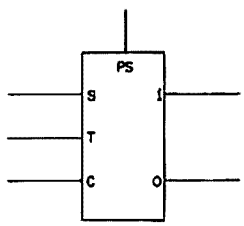
Let's first consider a gate. Its symbol, and a table describing its operation, are as follows:



A	B	OUTPUT
H	H	L
H	L	L
L	H	L
L	L	H

Note that H stands for a high-voltage level, and L stands for a low-voltage level. For the gate, the table shows a high level out for two low levels in, but low level out otherwise.

A flip-flop is represented by this symbol and truth table:



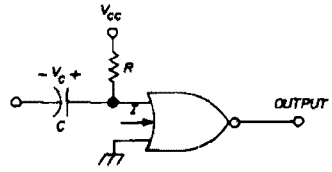
S	C	1 OUTPUT
H	H	NO CHANGE
H	L	H
L	H	L
L	L	COMPLEMENT

A high level on the preset (PS) terminal results in a low output at the 1 terminal. Any other change of state of the flip-flop requires a negative-going pulse on the toggle (T) input. (This pulse must be fast, I might add—less than 100 ns for the IC's used in this keyer.) The terminal-1 output after a toggle is shown in the table.

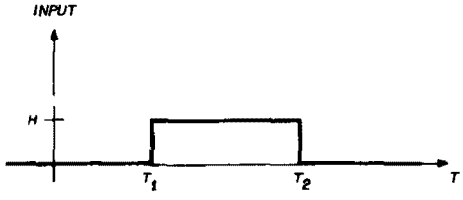
A printed-circuit board for this keyer is available for $3.00 from Stafford Electronics, 427 South Benbow Road, Greensboro, North Carolina 27401; order part no. HR 11-69A. A kit of IC's, transistors and diodes (less power supply) are available from the same source for $8.50.

With a low on S and C, the flip-flop changes state (complements), and the outputs return to the same state as the inputs after a toggle if there is an H-L or L-H combination on the inputs.

Now these are the only two different types of IC's used in the keyer, but three simple gate connections are described so a block diagram can be drawn. First, a monostable, or one-shot:



consider the following input signal:

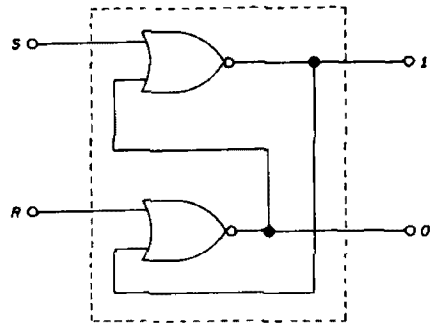


Before t_1 occurs, the capacitor is charged to $V = V_{cc} - RI$, where I is the current flowing into the gate, $V_{cc}/(R + 450)$. (The gate input resistance is 450 ohms.) The gate has an H on one input and an L on the other, so the output is L. At t_1, the input rises to H. The top input of the gate will rise to V + H, then the capacitor discharges to V − H. At t_2 the input falls to zero.

The voltage across the capacitor can't change instantaneously, so the top input to the gate is brought down to V − H, which will be a low level if H is close enough to V. The output of the gate rises to an H level. The capacitor then charges toward V through R_e.* When the capacitor voltage is greater than about 1 volt (a high level), the output falls back to a low. Hence, for an H-to-L transition **in**, we get a pulse of fixed duration **out.**

* R_e is an equivalent resistance that will be a "two-segment" resistor. It consists of R until the gate input transistor turns on, then of R in parallel with the gate resistance after turn-on. The voltage applied to the charge path is V_{cc} before the gate turns on, and V after it turns on.

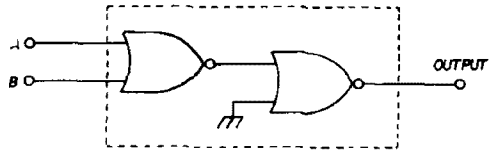
The second connection is a set-reset R-S flip-flop:



S	R	"1" OUTPUT
H	H	NOT ALLOWED (LOW)
H	L	L
L	H	H
L	L	USUAL CONDITION

For the flip-flop to change state, an alternate high level is required between inputs. Two high levels are not allowed. Even if you could remove them simultaneously, the resulting state would be indeterminate because of different IC switching times. However, during the time two high levels are applied, both outputs will be low, of course.

The third connection is an OR gate:



A	B	OUTPUT
H	H	H
H	L	H
L	H	H
L	L	L

The name implies an assignment of positive logic: H is designated 1, but the circuit is still characterized by the H-L description. (That is, a low output for two low inputs; a high output otherwise.) The second gate inverts the output of the first. Note that this circuit is an AND gate if an H level is designated zero and L is designated 1.

the TO keyer

The TO keyer is familiar to most amateurs, but for completeness, its block diagram is shown in **fig. 1A.** Clock pulses toggle FF1 whenever the clock is enabled. When

the dot contacts are closed, the clock and FF1 are enabled, producing the output as shown in **fig. 1B.** When the dash contacts are closed, both flip-flops are enabled, and the output looks like the waveform in **fig. 1C.** Hence we have a guaranteed one-to-one dot/space ratio and a three-to-one dash/dot ratio. This is the defining characteristic of the TO keyer.

Additions to the basic TO keyer block diagram for a dot memory are shown in **fig. 2.** When the dot contacts are closed, the memory is set and the clock and FF1 are enabled. This causes a dot to be formed. Once formed, a pulse from the output resets the memory, so only **one** dot is sent. With the block diagram in mind, let's look at the complete circuit (**fig. 3**).

toggling voltage

The clock is as described by K3CUW.[1] When a high level is at point A, Q4 is saturated. Because V_{CESAT} (saturation collector-emitter voltage) is less than V_{BE} (**ON** base-emitter voltage), Q5 is cut off. Point B is thus at 1.8 volts, C3 is charged to $1.8+V_{BE}$, and Q3 conducts the saturation collector current of Q4. When the dot or dash contact is closed, point A goes low, Q4 is cut off, and Q5 conducts. This reduces the voltage at point B, which biases Q3 even more into conduction, causing more base drive to Q5. The voltage at point B is further reduced. Finally, C3 is discharged to $V_{CESAT} + V_{BE}$, at which point Q5 ceases to conduct. The voltage at point B increases, which cuts off Q3 until the capacitor charges to $1.8 + V_{BE}$ again, then the whole process repeats. Result: nice, fast $1.8 - V_{CESAT}$ negative pulses—good for toggling flip-flops.

fig. 1. Basic block diagram of the TO keyer, A, with dot and dash waveforms, B and C.

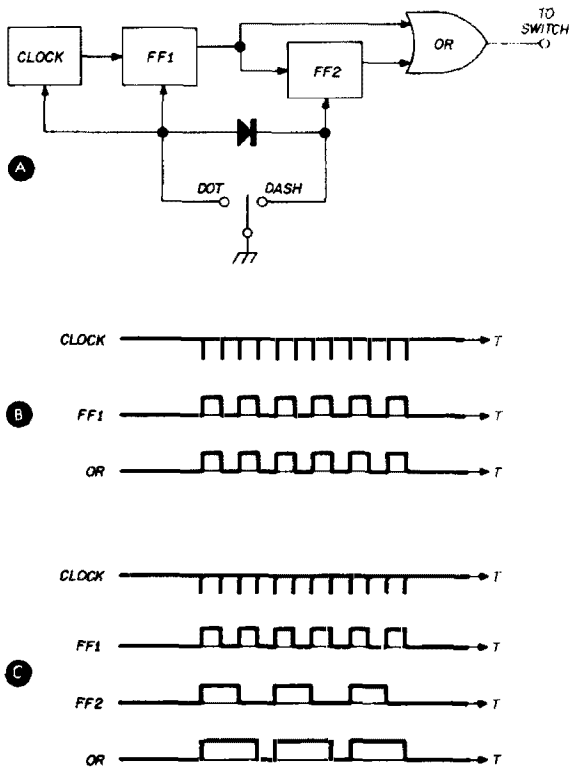
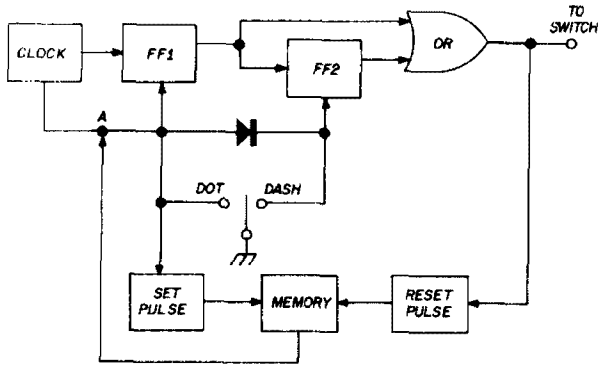


fig. 2. TO keyer with dot memory added.



The PS and S terminals of both flip-flops are grounded. There are only two possible input states for the flip-flop: complement (both S and C low) and a low output on terminal 1 (high input on C). A flip-flop can be considered disabled with a high input on C and enabled with a low input on C, if complementing is the desired operation.

A high input is applied to both flip-flops and to the clock (point A, **fig. 3**) when the keyer paddle is in the center, so everything is quiet.

dot formation

Closing the dot contact enables the clock, enables the dot flip-flop, and sets the dot memory. Recall that a high-to-low transition on the input of a one-shot gives a pulse out, and the terminal 1 output is low when a high voltage is on the top input terminal. Thus, point A is kept low by the memory when set.

The first clock pulse puts FF1 in a high

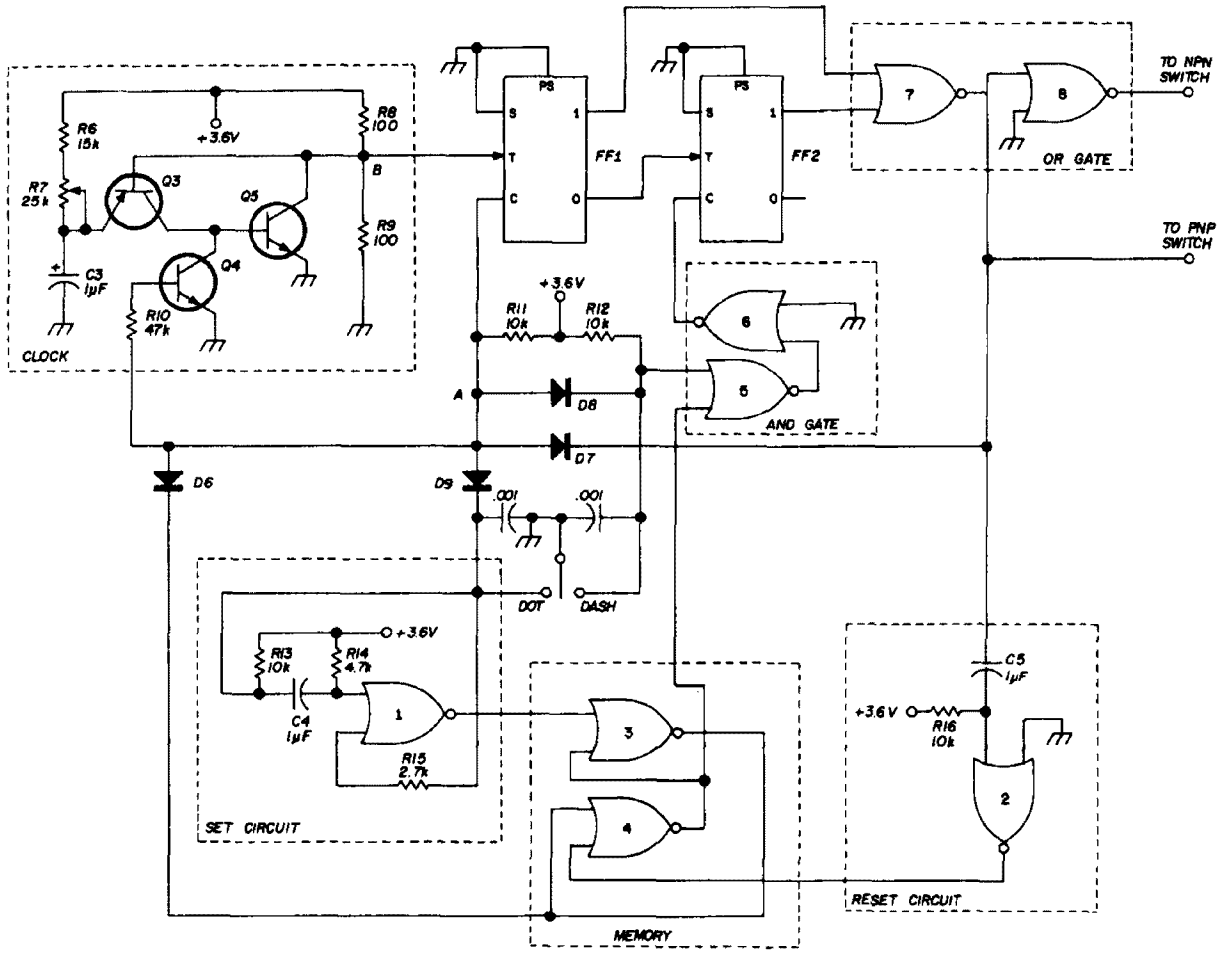


fig. 3. Complete schematic of the keyer. Memory and reset circuits ensure positive dot formation with no omissions.

state, and the high-to-low transition of FF1's 0 output puts FF2 in a low state, because its C terminal is high.

The second clock pulse puts FF1 to a low state. During this time, the clock remains enabled because point A is kept low by diode D8 (this makes the characters self-completing). Gate 7's output goes low as soon as the 1 output rises to a high on the first clock pulse. Also, the high-to-low transition at gate 7's output gives a reset pulse to the dot memory. The waveforms for a single dot are shown in **fig. 4.**

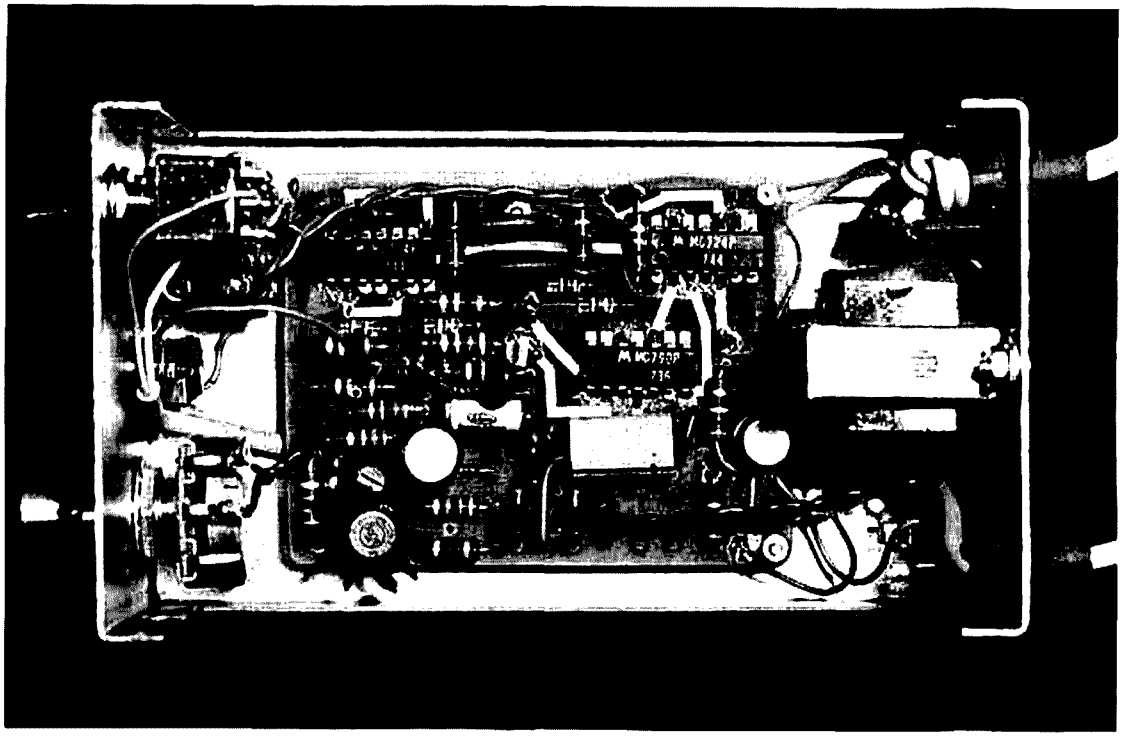
The reset one-shot provides a longer pulse than the set one-shot. Thus, when they start at the same time, the memory ends up reset. The reset pulse length is about 4 ms. Any shorter pulse would work as well, but 4 ms is long enough to ensure memory reset despite inevitable contact bounce from the paddle. A series of dots is produced by holding the dot contacts closed. The wave-forms in **fig. 4** are then merely repeated, except for the set pulse.

dash formation

Closing the dash contacts produces a series of dashes, as with the ordinary TO keyer. Note that FF2 is enabled or disabled by an AND gate. (To become enabled, the flip-flop requires a low input from the gate, which requires the dash contact output AND the memory output to be low.) The AND gate's function will be described later.

dash-dot sequence

Consider **fig. 5.** Even though the dot contacts were closed and released before the preceding dash was completed, the dot is generated. The low level from the memory



The completed keyer in a Minibox chassis. Pnp transistor switching is used.

keeps the clock and FF1 enabled until the dot is initiated; then the hold circuit, D8, finishes the dot—a neat, fail-safe arrangement.

the AND gate

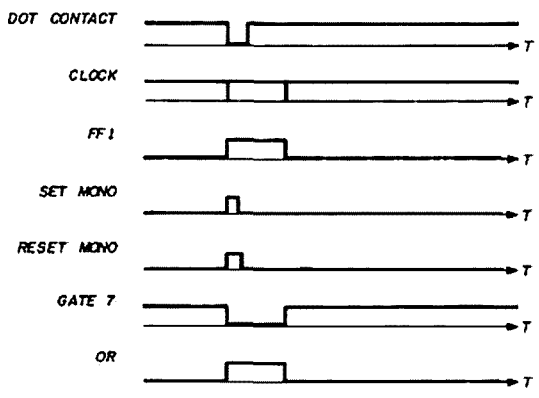
To illustrate AND gate operation, we'll consider forming the letter k. The waveforms in **fig. 6** result. The dot is remembered as before; however, the dash contacts are closed before the dot begins. If FF2 were not disabled, another dash would be initiated instead of a dot. The zero output of the memory is high whenever a dot is stored, so FF2 is disabled until after the dot is initiated. This feature allows "squeeze" operation, because the memory can be set even when the dash contacts are closed. Once the memory is set, a dot is formed following whatever character is currently being formed.

power supply

The effort to make the key small led to a search for alternatives to the large filter capacitors required in the conventional power supply[1] where "brute-force" filtering is used. The transistor-regulated power supply shown in **fig. 7** was chosen. The resulting size is smaller than one of the pair of filter capacitors.

Diodes D3-D6 form a bridge rectifier whose output is filtered by C1 and regulated with series regulator Q1. The reference voltage is the forward voltage drop of diodes D1 and D2 (about 1.2 volts, because they are silicon diodes). Transistor Q2 is the error-signal amplifier. Capacitors C1 and C2 were chosen by trial and error to reduce power supply ripple to less than ±10 percent, the tolerance in supply voltage specified for the

fig. 4. Waveforms for a single dot. Reset pulse length is 4 ms, which is long enough to ensure memory reset despite contact bounce.



IC's. Larger capacitors would, of course, reduce the ripple even more. R2 supplies base drive to Q1, and the regulation is obtained by Q2 controlling the current at the base of Q1.

R3 provides a few milliamperes of current to the reference diodes, putting them into a lower-resistance portion of their characteristic curve. The output voltage is determined by voltage divider R4, R5. Since the voltage at the base of Q2 is fixed at $V_{BE}+1.2$ volts, the voltage across the divider will adjust to give this value. That is, the output voltage is given by (assuming $I_B << I_{divider}$)

$$V_{out} = \frac{R4 + R5}{R5} (V_{BE} + V_{D1} + V_{D2})$$

The values shown are chosen so that the base current to Q2 (I_B) is very much smaller than the divider current ($I_{divider}$). Because individual diodes and transistors may have considerably different voltage drops, it may be necessary to adjust R4 to obtain the correct output voltage.

keyer circuits

The positive voltages from the IC's ordinarily require the switch to be an npn transistor. Such a keyer circuit is shown in **fig. 8A.** Notice that the key ground is elevated to the voltage across the key terminals of the transmitter. This arrangement requires a little care in mounting the components of the key in its box, and touching the key ground to the transmitter ground will short the transmitter key terminals.

A pnp transistor switch may be used but it requires a negative-going gate pulse to drive it. This gate signal is obtained from the output of gate 7, which is translated by offsetting the transmitter ground by the amplitude of the gate voltage. This offset is obtained by connecting the key supply to transmitter ground via three forward-biased diodes as shown in **fig. 8B.** R17 draws a little current through the diodes to ensure a uniform voltage drop with switching. Again, care is required in mounting key parts, and the consequence of shorting key ground to transmitter ground is more severe, because the key power supply is shorted.

Three of the four keys I've built have been sensitive to rf energy, to varying degrees. A 0.01 μF capacitor across the key power supply near the switch transistor seems to clear up the trouble, and a smaller capacitor often will do. The location of the capacitor may depend on the type of switch because of the grounding arrangement peculiar to the individual system. However, the larger

fig. 5. Waveforms for dash-dot sequence. Dot is generated even though dot contacts were closed and released before preceding dash was finished.

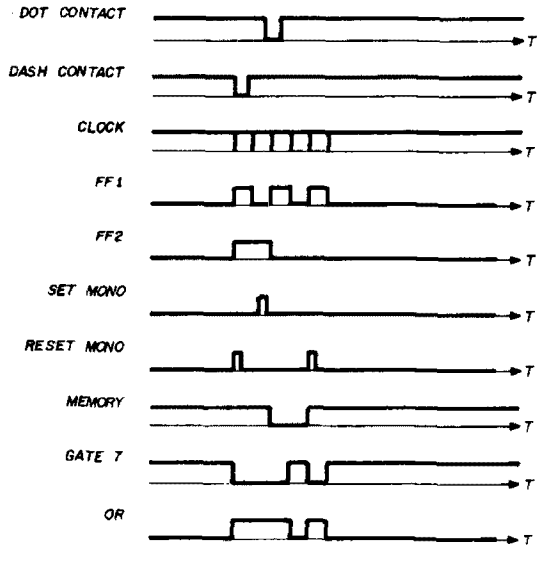
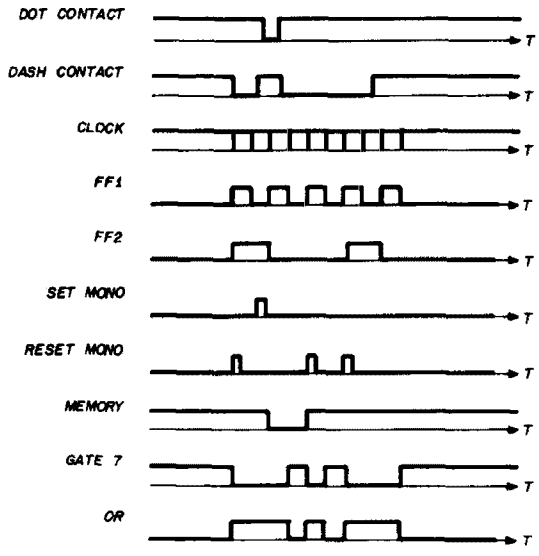


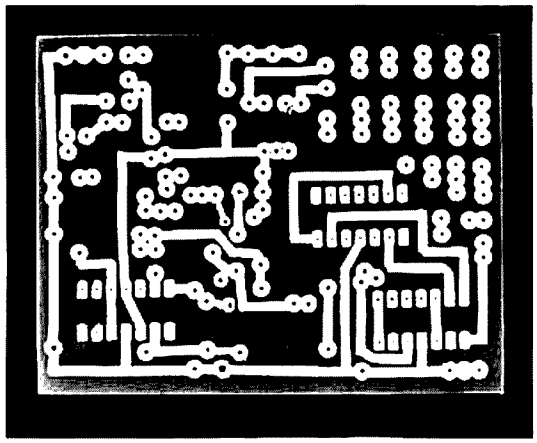
fig. 6. Waveforms illustrating the AND gate operation. This feature allows "squeeze" operation, because memory can be set even when dash contacts are closed.



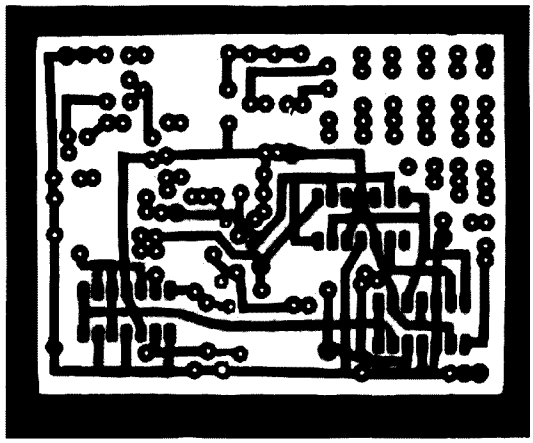
capacitor (0.01 μF is large here) seems to be noncritical as far as location is concerned.

construction

To make the key small, most of the components are mounted on a printed-circuit board. This board was designed to take ½-watt resistors, which is a lapse in the effort at miniaturization for the sake of availability of parts. However, ¼-watt resistors may be used for all except R1, which should be ½ watt. As the photos show, the board is quite uncluttered. The region of unconnected terminals is left to mount the switch transistor and its components. If the center row of these terminals is cut away, the board will fit onto the base of a Vibroplex bug between the pivot support and the terminals. I originally intended to mount the key on a Vibro-Keyer, but later decided on a minibox to leave room for additions such as a tone oscillator.



Bottom of the printed circuit board.



Printed circuit board viewed from below, top lighted to show top wiring.

choice of devices

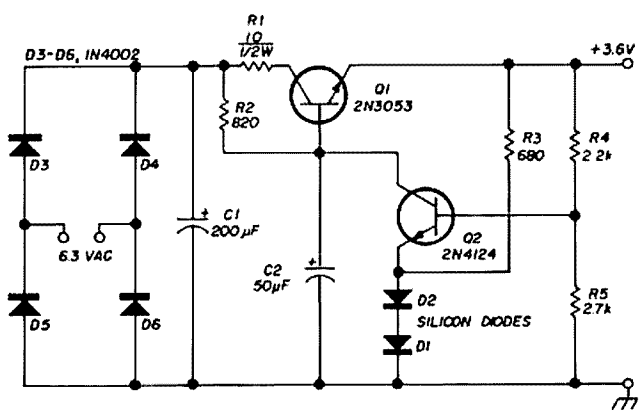
Motorola integrated circuits are used in the layout shown, but other RTL digital IC's are fine (the first keyer was breadboarded using Fairchild μL923 and μL914). The Motorola packaging has more devices with fewer leads, so they're smaller than the Fairchild IC's. The transistors are all noncritical, but they should be silicon. Q1, of course, must be capable of handling the power, and the switch transistor must handle the voltage and current at the key jack.

Mototola manufactures several suitable switches: MJE340 (300 volts), neon drivers 2N4409 (80 volts) and 2N4410 (120 volts), for example. The only pnp switch I am familiar with is the 2N398A (105 volts) manufactured by RCA, Sylvania, Motorola, and possibly others.

some recommendations

Capacitors C4 and C5 can't be electrolytic, because the polarity of the voltage across them changes. They must also be 1 μF or more to ensure that contact bounce doesn't cause the memory to be set. 1-μF, 3-volt ceramic discs (Centralab UK-105) are quite small and easily obtained. Diodes D6, D7, D8 and D9 must be germanium because of their lower voltage drop. Silicon diodes with their higher voltage drop cause some flip-

fig. 7. Regulated power supply for the keyer. R4's value may have to be adjusted because of different voltage drops in transistors and diodes.



flops to remain disabled even with point A low. However, silicon diodes are required for D1 and D2 because of their higher forward voltage and lower on resistance. The rectifier diodes are Motorola 1N4001 or 1N4002, because they're small. Any other diode with a rating of at least 100 mA and 25 PIV will work fine.

cured this problem by making R14 4.7 kilohm.

The keyer described here requires quite a bit of power: about 90 mA at 3.6 V. I built another keyer using Motorola milliwatt devices (MC776P flip-flops and MC717P gates) and replaced R8 and R9 with 470-ohm resistors. The new unit operates

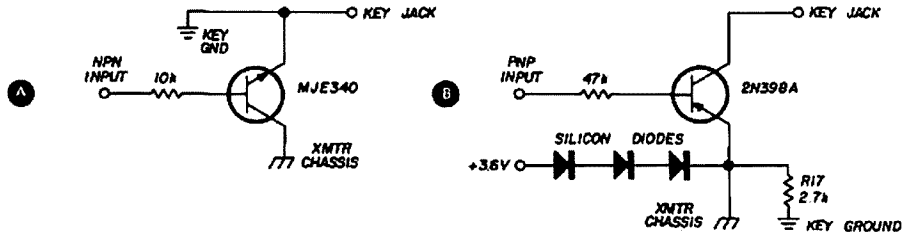


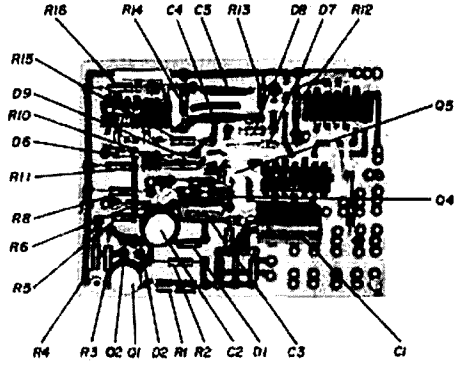
fig. 8. Keyer circuits. Npn transistor switch is shown in A; pnp switch in B. Care is required in mounting parts to avoid shorting key circuit grounds to transmitter ground.

The AND gate formed by gates 5 and 6 cannot be replaced with a diode AND, because the voltage drops are too high.

Resistor R15 was included in the set one-shot to reduce the sensitivity of this circuit to noise. If the input that's connected to R15 is returned to ground, as in the reset one-shot, the memory will occasionally be set when a noise pulse gets on the power-supply line. This will result in a stray dot.

Because electrolytic capacitor tolerances are so wide, several potential problems can develop. There may be a wide variation in the values of R6 and R7 in the clock speed control. Also, I originally used a 10 kilohm resistor for R14, which caused the set pulse to be longer than the reset pulse because C4 and C5 weren't close enough in value. I

Component layout.



from two penlight cells, and current drain is about 17 milliamperes!

A tone oscillator would be a useful addition. I'd appreciate hearing from anyone who comes up with a sine-wave oscillator-amplifier that will operate at 3.6 volts and drive a small speaker.

acknowledgements

Special thanks are due Otto Meier, HB9AFE, for laying out the PC board and assistance in designing the circuit. The photographs were taken by my brother, Richard Young and by Mark Hansen, VE7BGE.

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ham radio

antennas and capture area

Some basic definitions of the antenna as an energy transducer

Jim Kennedy, K6MIO, 5465 E. Hampton, Fresno, California 93727

It's almost impossible to discuss antennas without terms such as "capture area," "gain," "aperture," and "directivity" entering the conversation. Reason dictates there must be some interconnection between these concepts, yet incorrect statements are often heard such as "A colinear broadside is, in general, better for receiving than a Yagi since the colinear has more capture area." This indicates the relationship of these very important ideas is often subject to confusion.

There are many ways to analyze antenna performance. Some antennas, such as a dipole or an array of dipoles as in a colinear broadside, are often analyzed by considering (a) the net effect of the phase and amplitude of the fields from each individual dipole, and (b) the relative location of the dipoles. Such an analysis can become quite involved mathematically, but could be carried out for any array. With the aid of a computer you could produce a detailed plot of the antenna pattern in any desired plane and indicate the gain and other essential parameters.

While each type of antenna may have one convenient mathematical description, there are many ways to analyze any antenna. One general approach is to determine its **capture area,** or to use the precise language of physics, its **diffracting aperture.** In principle we can learn almost all we wish to know about an antenna's radiation characteristics if we know the dimensions of its capture area. It may seem surprising that this would produce the radiation field pattern, but it will, and other information as well.

Note I say "in principle." In practice the work to produce these results may be prohibitive, depending on the antenna. Determining the capture area for many antennas may be difficult, but adopting this approach will allow us to discuss any antennas qualitatively.

aperture

Since it forms the foundation of all that follows, the definition of capture area deserves some attention.

In most practical situations, one is concerned with how an antenna performs as a

radiator and how it performs with respect to a passing wave. The wave is sensibly plane if we consider a **small region** of space. That is, we may consider the wave as an endless succession of plane-wave fronts perpendicular to the line of travel.

The wave carries energy, so a certain power density is associated with it: so many watts per square meter of wavefront area. Every antenna has an area called the **capture area,** or **aperture,** that will intercept some part of the wave area. The capture area may be quite different than the area occupied by the actual antenna structure. This is discussed later.

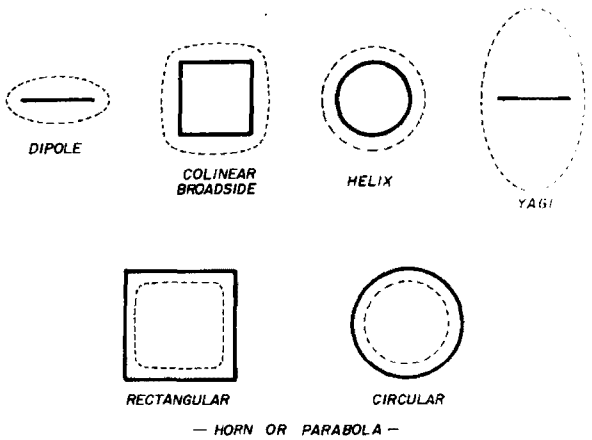


fig. 1. Typical apertures (dotted lines) for common antennas. Solid lines represent the boundaries of the physical structures.

We can now relate capture area to receiving antenna gain. If a wave front of five microwatts per square meter impinges on an antenna of one square meter capture area, five microwatts would appear at the antenna output terminals. A second antenna with twice the capture area would receive ten microwatts. Gain is therefore proportional to capture area.

The capture area of a half-wave dipole is usually used as a reference standard for measurements. Sometimes an **isotropic source** is used as a reference. Such a source radiates energy in all directions equally. (The sun is an example.) For an isotropic source, gain is given by

$$G = A \frac{12.6}{\lambda^2}$$

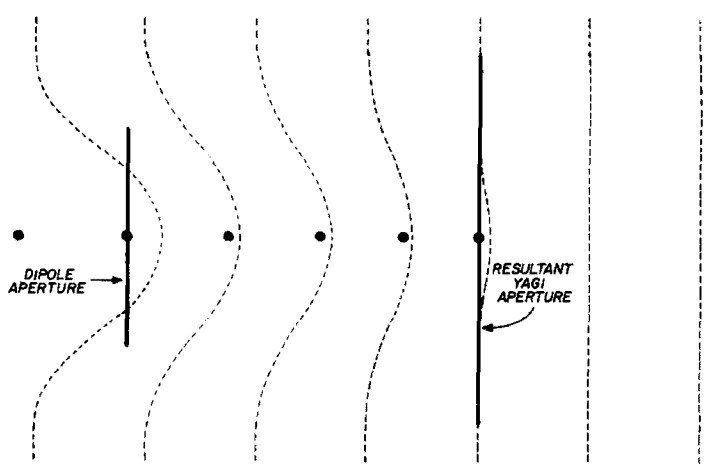
where A is the capture area and λ is the wavelength. The fraction in the right-hand term is a units conversion factor. The optimistic claims for antenna gain you sometimes see in advertisements are often based on an isotropic source. Obviously, any antenna characteristics based on this idealized reference should be viewed with caution.

Note that nothing has been said about the **shape** of the capture area. As far as gain is concerned, the capture area shape is superfluous. However, I'll explain later how the shape is important for other reasons.

size and shape

I mentioned that capture area may be different from the antenna structural area. A

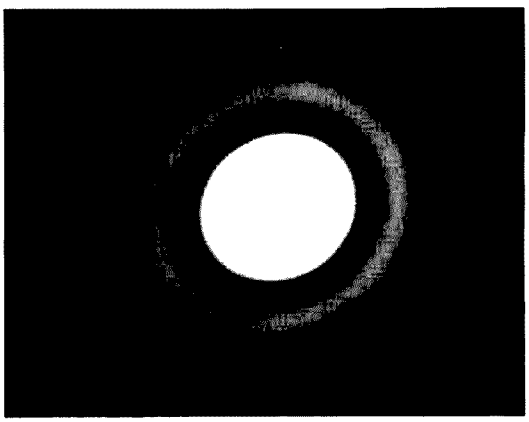
fig. 2. An edge view of a Yagi showing how plane-wave fronts (dotted lines) are bent towards the dipole by the director slow-wave structure. This results in a "magnification" of the dipole aperture.



simple example of this is the half-wave dipole. This antenna could be of very thin wire with extremely small physical area. It has a fairly large capture area, though, because it does deliver power to a receiver.

The dipole's shape implies an oblong aperture. Within limits, its aperture is independent of its physical area. A dipole's aperture is about 0.13 square wavelength. Comparisons of apertures of different antennas are given in **fig. 1.**

The colinear broadside antenna is an array of phased dipoles. Its capture area is somewhat larger than its physical structure.



This is a photograph of the diffraction pattern produced by a 16-wavelength circular aperture. The photograph was made by uniformity illuminating a 10-micron hole with red laser light. The light coming from the hole was then allowed to strike the film. The bright central spot represents the main lobe, which is about 8 degrees in diameter. The bright ring is the first side lobe.

This is because the capture areas of the dipoles overlap the total area. An analogy is a brick wall. Each brick (dipole capture area) has its own shape. In combination with the other bricks you see the entire wall, not each brick.

Another example of an antenna whose capture area is similar to its physical area is a dipole or waveguide horn with parabolic reflector. Both the colinear broadside and parabolic antennas are examples wherein aperture size and shape depend on physical structure. More details on the problems of parabolic antenna design appear in an earlier article.[1]

the Yagi antenna

One of the more difficult capture areas to analyze is that of the Yagi and helix antennas. Both are related, but the Yagi is more familiar to amateurs. Therefore, it will be discussed in some detail.

In the **plane** of the passing wave, the structural area of the Yagi is about the same as that of a dipole. Yet the Yagi has more receiving gain. Why?

The Yagi functions by a slow-wave structure composed of a series of directors (reflectors are not applicable here). As the wave passes each director, it excites a current in



This is another photograph of the 16-wavelength aperture pattern. The central portion has been deliberately overexposed to show the details of the much weaker side lobes.

the element. The director is shorter than a half wavelength, so these currents re-radiate a wave that is delayed in phase from that of the original wave. As the wave passes each director the wave front is slowed from its free-space velocity. The part of the wave farthest from the director is slowed less. As shown in **fig. 2** the wave folds in on itself. Thus, energy that would have missed the driven element is focused onto it.

Stated another way, the Yagi collects energy from a much larger area than the element itself. Therefore, the Yagi may have the same capture area and hence the same gain as a colinear array many times its physical size. This exposes the fallacy in the statement about colinears versus Yagis.

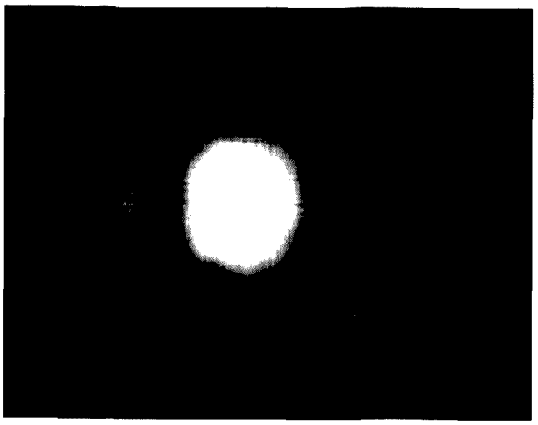
About the same can be said for the helix.

Here the wave is slowed because it tries to follow the helix, which means its forward progress is much slower than its speed along the spiralling wire.

The shape of the Yagi aperture is elliptical with its longest dimension perpendicular to the dipole. The circular helix has a circular capture area. (see **fig. 1.**)

diffraction

The preceding discussion established the existence of a real capture area of varying shape. While the gain has nothing to do with the shape of the aperture (only its area), the



This pattern was produced by a very poor square aperture. Note the more-or-less square shape of the main lobe and compare the side lobes with those of the circular aperture.

shape of the area does vitally affect the antenna **field pattern.** To understand this, it is first necessary to discuss diffraction.

Although the connection may not seem clear at first, consider a plane wave approaching a large sheet of conducting material with a hole in it, as shown in **fig. 3.** Further, assume the hole is **small** compared to a wavelength. Since the sheet is conducting, the wave can penetrate only through the hole. The electric field must satisfy the conditions that the component of the field parallel to the conducting surface, **at that surface,** must be zero. This condition is met on the left side of the sheet by the generation of a reflected wave (not shown), which is out of phase with the incoming wave **at the surface.**

On the right-hand side, this boundary condition can only be met if the field direction is perpendicular to the surface **at the surface,** since it is forced to start that way inside the hole.

The final result, as shown in **fig. 3,** is that the original wave, which was going in one direction, to the right, is now converted to a spherical wave spreading in all directions on the right-hand side of the hole. An observer to the right sees a plane wave coming from the hole and spreading into a spherical wave. In fact, the conducting sheet isn't even necessary. Whenever a plane wave emerges from a very small hole (in terms of wavelength), it spreads out. This is called **diffraction.**

To relate diffraction and **antenna pattern,** first observe that if an antenna receives energy from a certain capture area or hole, it must also **radiate energy from that same hole.**

antenna patterns

The principle of diffraction can be used to gain information about the antenna pattern

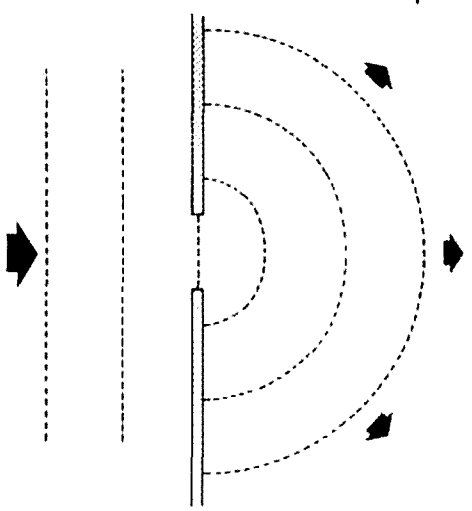


fig. 3. Diffraction of a plan wave from an extremely small aperture.

in general. Instead of working out the details, which require an application of calculus, I'll show the general approach and state part of the result.

Consider an aperture large compared to a wavelength. The aperture is made of a large number of very small, adjacent apertures (the brick analogy again). Each small aper-

ture will radiate more or less uniformly in all directions, as in the previous special case.

At some arbitrary distant point from the large aperture, an observer attempting to determine the net effect would find that each small aperture is out of phase with its neighbors. This is because each small aperture is at a slightly different distance from the remote point, depending on the angle with respect to that point. This is illustrated in **fig. 4** for the first and last "brick." Note that the distance to the first aperture is **a,** and to the last it is **a + b.** The radiation intensity at the remote point as a function of θ is the field pattern. To derive this, the mathematical calculation, in effect, shrinks all the small apertures to zero, then sums up this infinite number of apertures to find the total effect.

For our purposes it's unnecessary to give the complete expression, although the interested reader will find it in reference 2. Several important features of the pattern can be given. One is that there will be zero intensity, or nulls, at every angle such that

$$\sin\theta = n\frac{\lambda}{d}$$

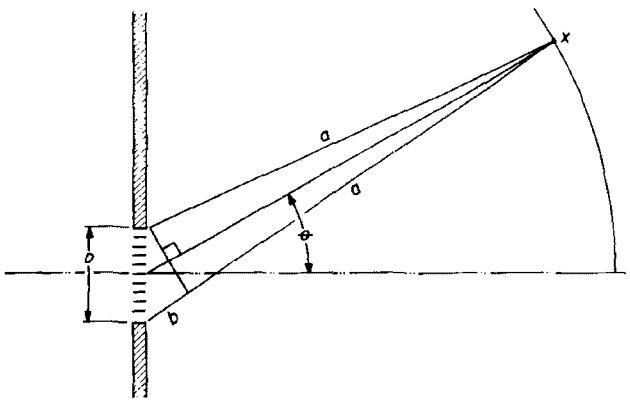
where λ is the wavelength, d is aperture length, and n can take on values of 1, 2, 3, . . . For each value of n there is some angle at which the pattern has a null.

If it is assumed that a peak exists between these nulls, it follows that the field pattern of a diffracting aperture is a series of peaks and nulls, which resembles an antenna pattern. **Fig 5A** illustrates this and includes all the usual details of antenna patterns.

Any two antennas, regardless of type, that have exactly the same capture area dimensions will produce the same field patterns at the same frequency. They would also have the same gain since they have the same area.

Another characteristic shows up in the null equation. The null angles are inversely proportional to the aperture dimension, hence the larger the aperture, the closer the nulls will be. That is, the larger the antenna, the narrower will be the beam.

fig. 4. Diagram for the calculation of diffraction patterns in the general case.



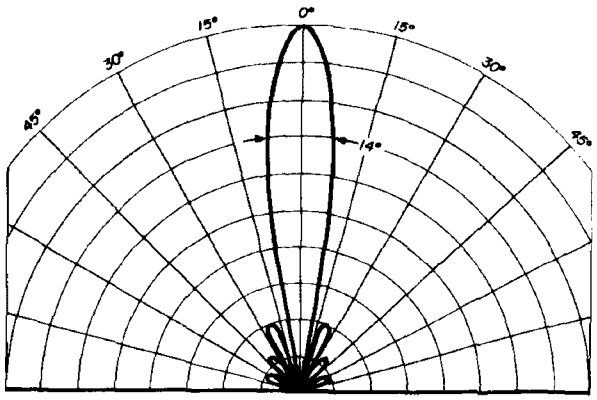


fig. 5. Typical diffraction patterns.
(A) Uniformly illuminated aperture four wavelengths in diameter (above). (B) The same pattern, but assuming a rear aperture of the same size attenuated 20 dB.

stacking

Only one aperture dimension was taken into account in the preceding derivation. As a result the derived pattern is for only one plane. The horizontal and vertical patterns are functions of the aperture horizontal and vertical lengths. Hence the two patterns are independent. What happens if a Yagi system is stacked?

Two Yagis stacked one above the other have a vertical beam twice as narrow as a single Yagi. The opposite is true for side-by-side stacking. Four antennas in an H-array would have one-half the beam-width in both planes, and so forth.

In the same way the main lobe is modified by changes in the aperture shape, sig-

nificant effects are produced in the side lobes. As the main beam is compressed by stacking, the side lobe's locations are altered. Old side lobes will disappear and new ones will appear elsewhere. While the angular beam dimensions on the horizontal and vertical axes are independent, off the axes they are not. That is, both the main beam and side lobes off the two main axes are dependent on other dimensions of the aperture.

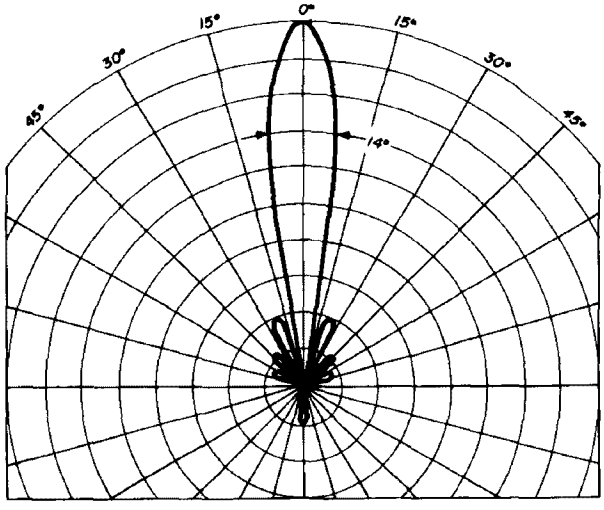


fig. 5B.

transmitting and receiving antenna gain

I showed previously that receiving antenna gain is a function of capture area. Transmitting antenna gain is a function of antenna beamwidth. The smaller the beam, the greater will be the forward field. All this boils down to a simple fact: transmitting antenna gain is directly proportional to capture area, as is receiving antenna gain. There is nothing new about this. Way back in 1929 the relationship between antennas as receivers and transmitters of electromagnetic energy was precisely stated in the ***reciprocity theorem,***[3] which is valid for any antenna. This merely says that any antenna that has gain as a receiving antenna will have the same gain as a transmitting antenna, **providing** the nature of the wave remains unchanged as it propagates between receiving and transmitting antennas.

the dipole

How does this simple antenna generate its capture area? The answer is not obvious. In some respects, the dipole is almost too simple to use the concept of an aperture. The main problem in trying to identify its shape is that in one plane its pattern is omnidirectional. If you insist, an aperture in the shape of a cylindrical sheet wrapped around the dipole might suffice, but this is at variance with the flat-surface concept used in the more sophisticated cases. In most arrays of interest, the flat aperture is a good approximation of reality.

back lobes

When back lobes appear, it is because the antenna has a radiating aperture to the rear as well as to the front. A colinear array without reflectors, for example, will radiate equally well in both directions. The forward and rear field patterns will be identical. A reflector will attenuate rearward radiation by its shielding effect, **fig. 5B.** It is possible in some cases to eliminate sidelobes entirely.[4]

conclusions

I have shown that an antenna has the same pattern whether used as a transmitting or receiving antenna. I have also made some simplifications to explain the basic principles without rigorous mathematical treatment. If you are interested in pursuing the extensions of the analysis of the two-dimensional aperture approach, a detailed discussion is given in reference 5.

The next time you become involved in a heated discussion on antennas, you'll now have some ammunition when statements such as those at the beginning of the article are made.

references

1. Jim Kennedy, K6MIO, "Illumination and Parabolic Antenna Design," *73 Magazine*, December, 1966, p. 90.
2. F. A. Jenkins and H. E. White, "Fundamentals of Optics," McGraw-Hill, New York, Chap. 15.
3. R. Carson, "Reciprocal Theorem in Radio Communications," *Proc. I.R.E.*, June, 1929, Vol. 17, p. 952.
4. J. D. Kraus, "Antennas," McGraw-Hill, New York, 1950, p. 93.
5. A. Z. Fradin, "Microwave Antennas," Pergamon Press, Long Island, New York, 1961, Chap. 3.

ham radio

increased sideband suppression for the HT-37

Easy conversion to a filter-type sideband generator

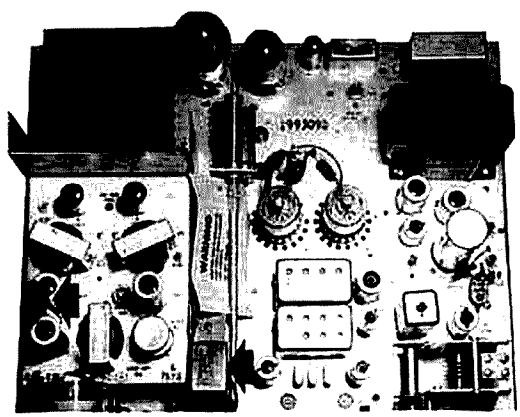
Milton L. Pokress, W3CM, 68 Sheridan Street, N.E., Washington, D. C. 20011

If you're an owner of a Hallicrafters HT-37 ssb transmitter, I'm sure you'll agree it has many excellent features. One area where it could be improved is in its sideband generator. The HT-37 uses the phase-shift system. It's difficult to obtain really good attenuation of the unwanted sideband with this method. The phase-shift networks are tricky to adjust and to keep adjusted. The unwanted sideband in most phasing systems is down about 30 dB, while 45 to 55 dB is not uncommon for filter systems.

The HT-37 can be modified to use a filter system for about $35.00. You should have no trouble adapting the circuit to the filter system described in the following paragraphs. You'll need the parts in table **1.**

filter selection

After examining many available filters, I chose the McCoy Silver Sentinel. I paid



Top of the modified HT-37 chassis. Left arrow shows position of extension shaft coming up between tubes V1 and V2. Knob is shown for illustration and is installed after top cover is in place. Filter (right arrow) is mounted on a bracket and attached to one of the holes that hold the subassembly in place.

$32.95 for mine direct from the manufacturer.* Its features are listed in **table 2.** I chose this filter because its input impedance closely matched that of the circuit and didn't require any matching network. Also, its low insertion loss eliminates the need for an amplifier ahead of the mixer.

installing the filter

A check of your HT-37 schematic will show that the carrier oscillator is one-half of a 12AT7 (V2B). If you look at the top of the chassis you'll see this tube, the audio stages,

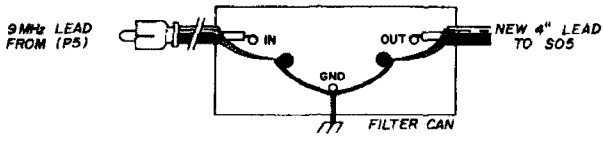


fig. 1. Underside of filter. A phono jack is soldered to the IN pin and a 4" piece of shielded lead soldered to the OUT pin with a new phono plug as shown.

balanced modulator and phasing network on a subassembly located on the left-hand side.

A shielded lead (P5) comes out of the front top of the subassembly chassis and goes to a phono jack (SO5) on top of the main chassis near the long shaft that turns the driver tuning capacitor. Pull P5 lead out and wire in the filter as shown in **fig. 1.**

* McCoy Electronics Co., Mt. Holly Springs, Pennsylvania 17065; price includes two sideband crystals. Suitable 9-MHz crystal filters are also available from Spectrum International, Box 87, Topsfield, Massachusetts 01983. The XF-9A filter, at $19.95 is a 5-pole unit; the XF-9B, an 8-pole filter (better shape factor), is $27.50; matching sideband crystals are $2.50 each.

Table I. Parts list for the HT-37 modification.

quantity	part
1	3-position rotary switch and pointer knob
1	aluminum stock for switch bracket
1	bracket for filter mount (see photo)
1	1/4- to 1/4-inch shaft coupler
1	1/4-inch shaft extension (see text)
1	McCoy crystal filter (see text)
2	25-pF trimmers
2	crystal sockets
1	phono slug and jack

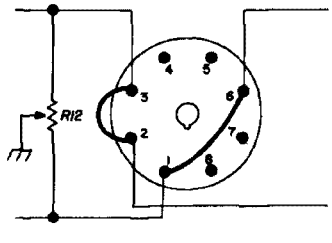
crystal switching

Before starting to work on the underside of the chassis, pull out Z1, the 90-degree phase-shift network located next to T101. It's in an octal socket and comes out just like a tube. Short pins 2 and 3, and 1 and 6 with bare hook-up wire as shown in **fig. 2A,** then plug it back in. The audio phasing network is now disabled.

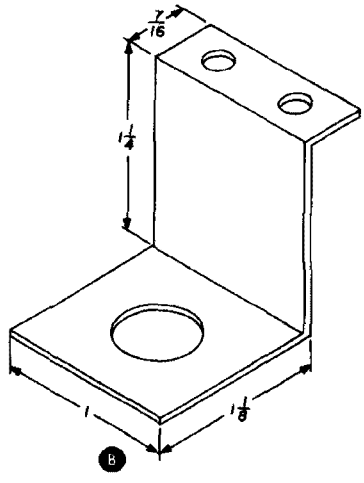
On the underside of the subassembly chassis, modify the carrier oscillator circuit as shown in **fig. 3.** Break the lead at point X and wire it to the pole of a 3-position rotary switch. Mount the switch on a bracket close to the grid of V2B.

Drill a hole large enough to pass the shaft of the rotary switch exactly between the center of V1 and V2. Extend the switch shaft with a coupler and a brass or aluminum extension shaft. Make the extension long enough so about one-half inch will fit through the cover of the transmitter when it is back on. File or turn down the diameter of the top end of this shaft sufficiently so it will

fig. 2. New connections to the phase-shift network are shown in A. Crystal switch bracket is shown in B.



A



B

fit through one of the large perforations on the top cover. The knob, of course, will be put on after the cover is in place.

With the wiring completed and the two new crystals installed, your switch should select the 8998.5-kHz crystal, 9-MHz crystal, and 9001.5-kHz crystal in that order. Now, snip L102 lead where it connects to C116; also snip RFOC capacitor lead where it connects to C117. These are green wires. The modification is now complete. The phasing networks are disconnected, and the filter, with its corresponding crystals to produce upper or lower sideband, is ready for tune-up.

Since the frequency of the 9-MHz crystal falls right in the center of the filter passband,

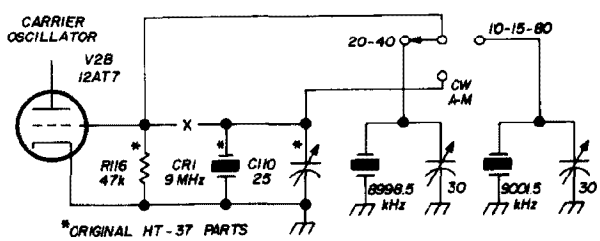


fig. 3. Modification to the carrier oscillator circuit. The lead from the grid is broken at x and connected to the arm of the three-position switch. The 9-MHz crystal and its associated trimmer are part of the original HT-37.

cw and a-m operation is undisturbed. When you operate in these modes, the new switch must be in the 9-MHz position. The upper and lower sideband positions on the main function switch on the front panel are inoperative, and either upper or lower sideband is selected by the new switch. However, the function switch must be left on upper or lower sideband simply to get it off the cw or a-m positions when operating sideband.

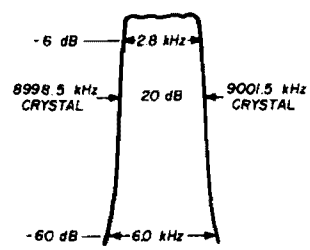
Table 2. McCoy ssb filter characteristics.

center frequency (MHz)	**9.0**
bandwidth (Hz)	**2360**
shape factor	**2**
sideband attenuation (dB)	**45**
insertion loss (dB)	**2.3**
input impedance (ohms)	**660**
crystal frequencies (kHz)	**8998.5 and 9001.5**

adjusting the filter slope

The manufacturer advises that the frequency of the 8998.5 and 9001.5-kHz crystals should be adjusted to fall 20 dB down the slopes of the filter. In selecting these points, the filter will work at its optimum design characteristics, pass the desired sideband fully, and suppress the unwanted sideband

fig. 4. Characteristic curve of the McCoy 9-MHz crystal filter.



most effectively. **Fig. 4** shows what the characteristic curve looks like and will give you an idea how we will find the 20-dB point on the filter slope. You don't need any fancy instruments, but you must have a vtvm with an rf probe.

After a 30-minute warm up, tune up the transmitter at 3800 kHz using the 9-MHz crystal for maximum output into a dummy load. Rebalance for maximum carrier suppression. Set your vtvm on the 100-volt scale and for minus dc. Connect your rf probe with an alligator clip to the transmitter's antenna output terminal inside the chassis. Ground the ground lead of your probe to the chassis. Select the 8998.5-kHz crystal, set the function switch on either upper or lower sideband, and set the audio control to zero. Turn the operation switch to MOX and unbalance the left carrier balance control, and at the same time, using an insulated tool, (this is essential) turn the trimmer capacitor corresponding to the 8998.5-kHz crystal for maximum output on the vtvm.

You will note, as you change the frequency by turning the trimmer, the reading will go up then level off. What is happening is that at the maximum reading, you had the crystal frequency on the top of the filter as shown in **fig. 4.** Note your maximum read-

Table 3. Crystal selection for desired sideband.

amateur band	sideband	crystal (kHz)
10 & 15	upper	9001.5
20	upper	8998.5
40	lower	8998.5
80	lower	9001.5

ing—mine read 65 volts. Whatever your maximum reading is, you must back it off to read only 10 percent of this. In my case, since 65 volts was maximum, I backed it down to 6.5 volts (change your vtvm scale for accuracy when you get the reading low

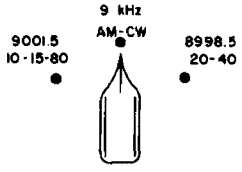


fig. **5. Label switch as shown. With this arrangement, lower sideband is selected for 40 and 80 meters and upper sideband for 10, 15 and 20.**

enough). You are now 20 dB down the slope of the filter. Repeat this procedure with the 9001.5-kHz crystal. Now switch to the 9-MHz crystal and rebalance the carrier for maximum suppression. You will note that when you switch to either of the two sideband crystals, carrier suppression will be even better.

selecting the proper sideband

A process of up conversion and/or down conversion is used in the HT-37 to arrive at the operating frequency on the various ham bands. **Table 3** will assist you in selecting the proper crystal to operate on the customary sideband. Label your new selector switch as shown in **fig. 5** to correspond to the sideband used on a particular band.

No provision has been made to compensate the vfo automatically for the difference in frequency between the upper- and lower-frequency crystals if you were to switch from upper to lower sideband or vice versa on any one band. It would get rather complicated on this transmitter as the vfo frequency would have to be increased on some bands and decreased on others. It's not worth the trouble since one doesn't keep switching sidebands when working on a particular band. If you wish to change sidebands during a contact, you can shift the main tuning manually to remain on the same frequency.

Before putting the top cover back on, bracket the filter to one of the existing threaded holes that holds the subassembly shield in place as shown in the photo. In this position the filter is in the clear and away from any heat generated by the tubes.

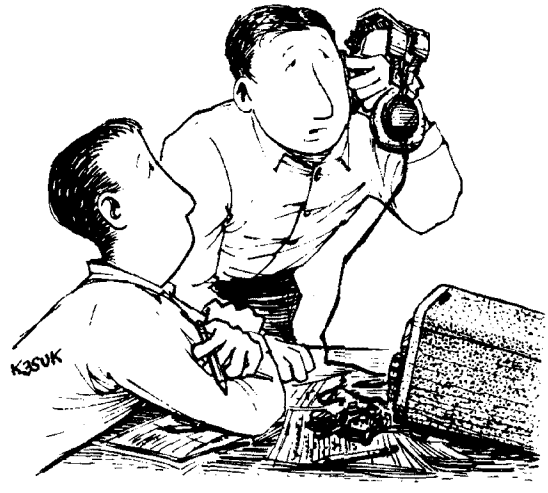
final checks

As a final check, try contacting some of your local buddies and ask them to check your quality and sideband suppression. Your unwanted sideband suppression should be at least 45 dB down. They may say your voice sounds a little different. The filter is quite sharp, and the voice frequencies passing through the filter bandpass will account for the difference. In any event you should now have a cleaner, sharper signal with practically all your power concentrated in the wanted sideband.

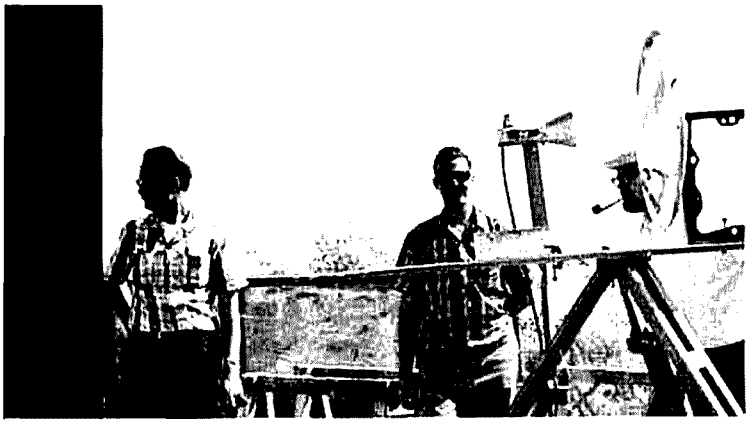
A word of caution—the HT-37 always had plenty of audio drive; now you have a bit more. It's very easy to flat top, and one of your best investments would be a monitor scope, if you don't already have one, so you can adjust your drive "right on the nose."

Congratulations—you are now the owner of a solidly built, modern filter-type single sideband transmitter.

ham radio



"No wonder you're having trouble with the code—that's RTTY!"



The "Saucer Sled" antenna is being bore-sighted by K6HIJ on a home-brew antenna range. W6DSL and WA6HWV look on.

a low-cost amateur microwave antenna

Richard B. Kolbly, K6HIJ, 26335 Community Boulevard, Barstow, California 92311

One of the major problems with putting an amateur microwave station together is obtaining a suitable antenna. For work above 2300 MHz, amateurs have traditionally used parabolic reflectors which they have either located on the surplus market or built themselves. It takes a fair amount of luck to find a surplus dish at a reasonable price, even if one is fortunate enough to live in an area that has a number of surplus dealers. Making your own dish is anything but easy, especially if you want it to be effective at 10,000 MHz.

raw materials

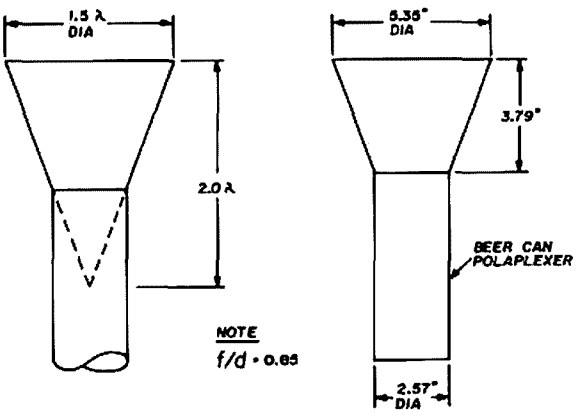
In the 1968 Sears Christmas catalogue I noticed a "Saucer Sled."* It was 26 inches in diameter and made of "sturdy aluminum." It looked like a natural for a dish, and in the interest of science, one was ordered. When it arrived, inspection showed it was spherical, with a usable surface 24 inches in diameter. It was indeed well built, and would make an excellent antenna for amateur microwave work.

The focal length was measured by focusing the sun's rays on a card (**DON'T** try using your hand as a focal plane!) and was approximately 15½ inches. Since this corresponds to a f/d ratio of 0.65, instead of 0.4 for a typical parabolic reflector, a standard open-end waveguide feed such as a polaplexer[1] will not properly illuminate the reflector surface.

construction

A conical horn feed was designed from the literature[2] that would have the proper beamwidths to illuminate the "Saucer Sled"

fig. 1. Feed system dimensions at 3300 MHz for the "Saucer Sled" antenna. The polaplexer is a half-quart empty beer can.



effectively. The feed was constructed with flashing copper attached to a homemade polaplexer (**fig. 1**) for the 3300-MHz band. The horn alone will produce a gain of 12 dB over an isotropic source.

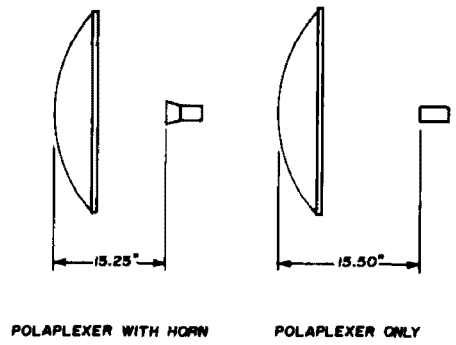


fig. 2. Measured feed locations for best gain.

The gain of the antenna, measured at 3335 MHz, was 24 dB (within the accuracy of measurement). If the reflector is illuminated with a simple polaplexer instead of the horn, the gain will be 1 dB less. These values agree closely with theory and are well within the errors of measurement. **Fig. 2** gives the location of the feed for best focus in both cases.

* This "Saucer Sled' microwave antenna is a simple and effective solution to the antenna problem for microwave work. I hope this will eliminate one stumbling block for the amateur who wants to experiment with microwave communications. I'd like to thank W6DSL and WA6HWV for their help in evaluating the performance of this antenna.

You should check the focal length of your own "Saucer Sled," as my measurements were made with only one sample, and I don't know how closely the dimensions are controlled during manufacture. If you can find a store that stocks these "Saucer Sleds," select one with a minimum of dents. Mine arrived with some dents that probably occured during shipping. Unless the dents are quite severe, however, they should not appreciably affect performance.

the mount

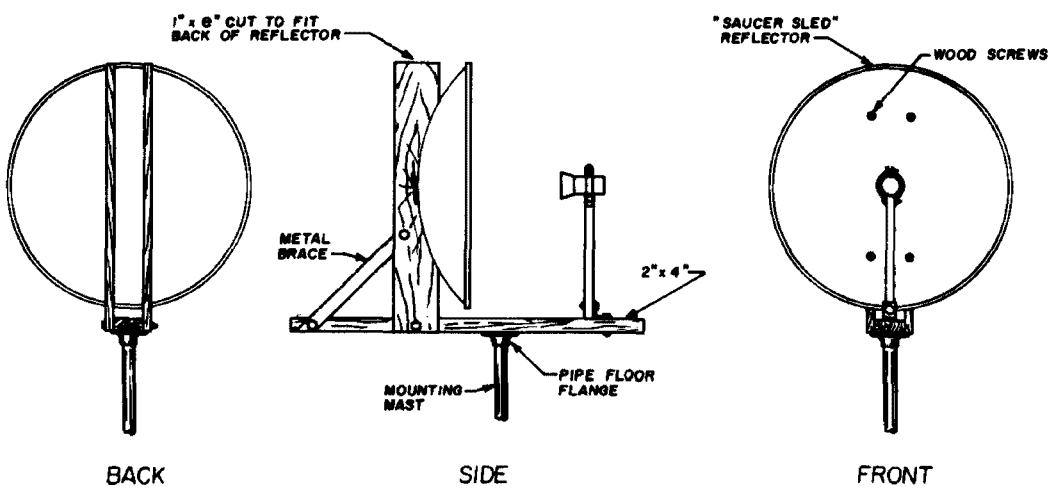
A suggested mounting method is given in **fig. 3,** but feel free to use your own ideas. Bear in mind there should be no relative movement between reflector and feed, nor between antenna and mount. Be sure to allow for focusing adjustments and elevation angle, because this antenna has a narrow beamwidth: about 10 degrees at 3300 MHz. The beamwidth is proportional to frequency (i.e., about 3 degrees at 10,000 MHz.) The feed should have a total of two wavelengths (approximately) of travel around the measured focal point along the antenna axis. This is 6 inches or so at 3300 MHz.

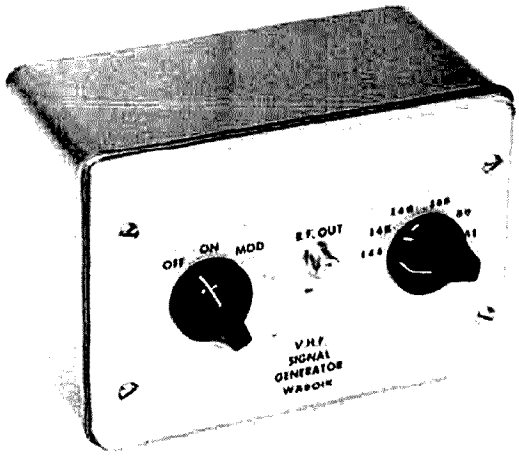
references

1. K. E. Peterson, K3KRU, "Practical Gear for Amateur Microwave Communications," *QST*, June, 1963, pp. 17-20.
2. G. C. Southworth, "Principles and Applications of Waveguide Transmission," Van Nostrand, 1950, pp. 415-9.

ham radio

fig. 3. A suggested scheme for mounting. Rigidity is important because of the narrow beamwidth.





a tone-modulated signal generator for two and six meters

Eliminate the guesswork in alignment problems with this compact vhf signal source

Vernon Fitzpatrick, WA8OIK, McLain Park, M 203, Hancock, Michigan 49930

After several years of trying to align home-built vhf converters and converted surplus equipment using harmonics from a cheap kit-type signal generator, I decided it was time for a change. Commercially built and surplus vhf generators were beyond my budget. I had thought seriously about building an 8-MHz harmonic generator, but the problem with these is knowing for sure you're aligning on the correct harmonic.

I started the project described here with the idea of building a temporary signal source for two meters. I ordered an OX-H oscillator kit with an EX crystal ground for 48.000 MHz.* The crystal's third harmonic was to provide the two-meter output.

The oscillator was much better than I expected. It has a range of 45 to 60 MHz without retuning. The specification sheet with the OX-H oscillator kit stated that an AOS-A1 audio oscillator could be used to tone-modulate the crystal oscillator. So, here were the major elements for a stable, accurate, modulated signal generator. I then ordered the audio oscillator, and the temporary signal source evolved into a permanent item of

test equipment with all the features I needed (see **fig. 1**). A crystal switch would provide signals for six and two meters without retuning the oscillator. The crystal frequency could be used for 50 to 54 MHz, and the third harmonic of crystals in the 48.000 to 49.000 to 49.333 MHz range could be used for output on 144 to 148 MHz.

I have six marker frequencies and can add five more. Most two-meter activity is on the low end of the band, so I have more marker signals in this region. For two meters, my marker signals are 144, 145, 146, and 148 MHz. I also plan to get on six meters, so my signal generator has outputs on 50 and 51 MHz.

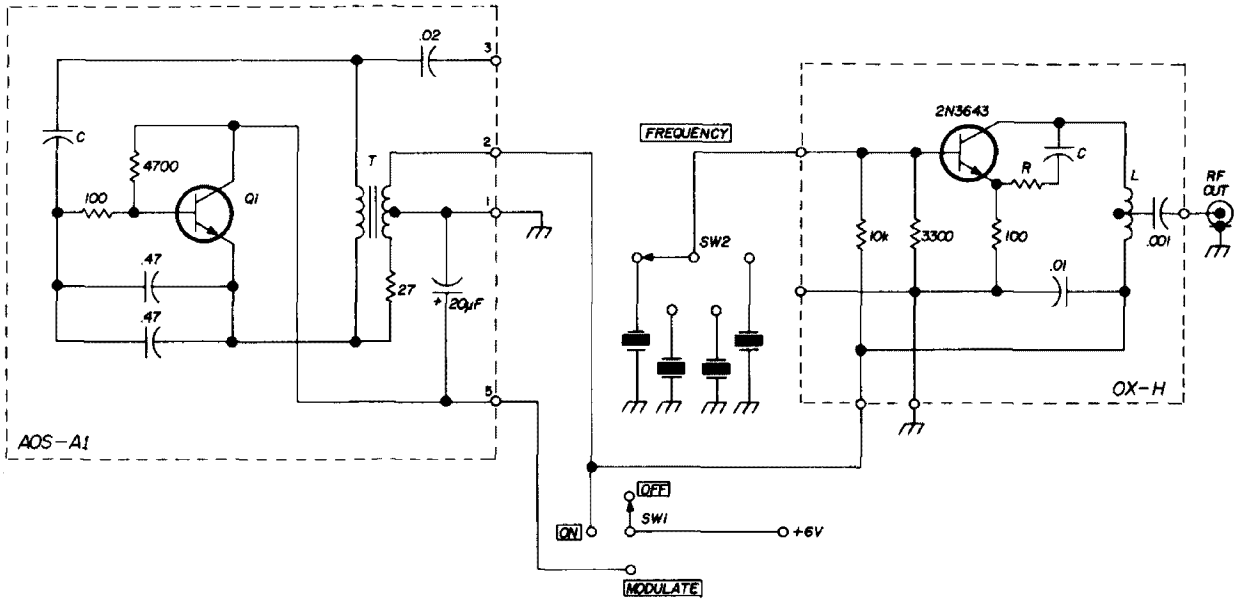


fig. 1. Schematic for the vhf signal generator. Four crystals are shown; up to eleven can be used.

construction

I used a Bud CU-463 cabinet, because I happened to have one in the junkbox. However, the chassis will fit into a 6x6x6-inch cabinet; for ease of inserting crystals, I recommend the latter.

Assemble the OX-H oscillator according to the instructions with the kit. Touch-solder the crystal socket so it can be removed after testing. Check out the oscillator by listening for the signal in a receiver. When the oscillator is working properly, remove and discard the crystal socket. Set the oscillator aside for the time being.

The AOS-A1 audio oscillator has one mounting hole at the rear. The audio oscillator must be insulated from the chassis. Use a sharp knife to remove the copper foil for approximately 1/8 inch around the mounting hole. To provide two-hole mounting, I drilled a 9/64-inch hole in the clear area behind terminal 6. This completes the preparation of the modulator assembly.

crystal switch

A Centralab PA1001 phenolic switch was used for the crystal switch assembly. The phenolic switch has a shorter and more direct path than ceramic switches.

Prepare the crystal sockets by bending one terminal out. Solder the bent terminals directly to the switch terminals. Crystal sockets 1, 5 and 6 should be almost vertical. If eleven sockets will be used don't solder sockets 10 and 11 until the audio oscillator is mounted on the chassis. Posi-

* International Crystal Mfg. Co., Inc., 10 North Lee, Oklahoma City, Oklahoma 73102. (OX-H kit, $2.95 ppd. EX crystals, $3.95 each ppd. AC-5 crystal socket $.15 each, shipping weight 2 ounces. AOS-A1 transistor audio oscillator, $9.95, shipping weight 2 pounds.)

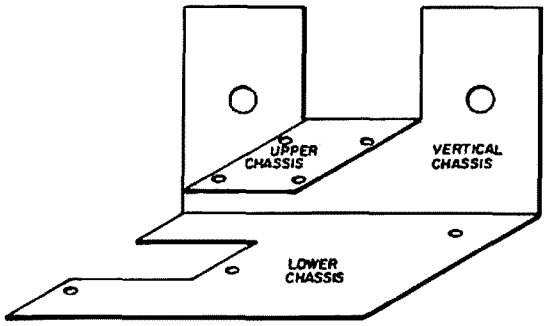


fig. 2. Layout for the chassis. A piece of aluminum is used, measuring 4¾ by 6 inches.

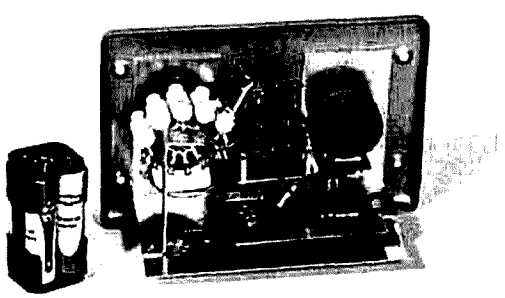
tion crystals 10 and 11 so they'll clear the AOS board. Remove stop tabs on the switch to correspond with the number of crystal sockets used.

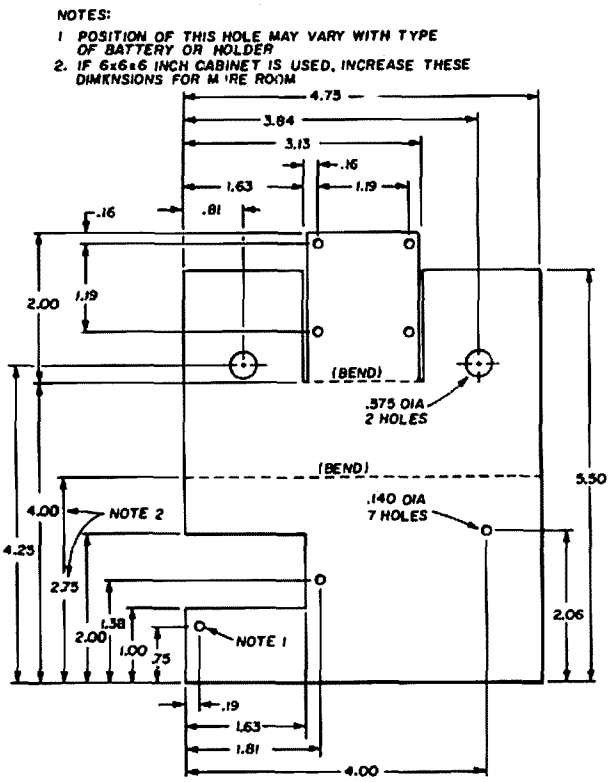
Make a 1-3/8-inch diameter circle of number 16 bare or tinned copper wire, and leave a 1-inch pigtail. Place the wire circle behind the crystal switch so the pigtail is lined up with the switch center contact terminal. Solder the ground terminals of the crystal sockets to the wire circle, then solder a 1-inch pigtail of number 16 wire to the switch center contact. Set the switch assembly aside.

the chassis

Lay out the chassis on a 4-3/4x6-inch piece of aluminum in accordance with the dimensions given in **fig. 2.** Bend the bottom chassis back. Position the audio oscillator with terminals flush with the left (from the front) edge and centered from front to rear, then mark and drill the mounting holes with a 9/64-inch drill.

Bend back the upper chassis, and drill four 9/64-inch holes for the rf oscillator. (The rf oscillator will be positioned flush





with the rear of the chassis.) Next, drill the two switch holes on the vertical chassis. The cabinet front panel is drilled to correspond with the switch holes in the vertical chassis. Depending on the type of jack you use, drill the rf output jack hole between the switch holes, from ¼ to ½ inch above. Be sure the jack clears the rf chassis.

final assembly

Begin assembly by mounting the rf jack on the cabinet front. Mount the rf oscillator on the chassis, using the mounting hardware supplied with the kit. Be sure the rf terminals are to the front. Secure the chassis to the front panel with the switches. Pass the switch through the chassis hole, then through the front panel hole, and tighten the mounting nut. Mount the crystal switch so the pigtails are as close as possible to the crystal holes in the OX board. Pass the wires through the holes, cut off excess, and solder.

The function switch (SW 1 in **fig. 1**) carries only dc, so any three-position switch can be used. Using the terminals supplied with the OX kit, wire the rf and ground to the rf output jack. Mount the AOS board

table 1. Specifications for the vhf signal generator.

frequency range	45 to 60 MHz and harmonics
frequency tolerance	0.02 percent of crystal frequency
operation temperature range	0 to 50 degrees centigrade
rf output at crystal frequency	0.2 V rms into 50-ohm load
modulation frequency	1000 Hz ±10 percent
audio distortion	less than 15 percent
power requirements	6 Vdc unmodulated at 20 mA; modulated, 70 mA

using two 3/8-inch long insulated spacers. Connect the battery positive lead to the rotary terminal of the function switch. Connect a wire from number 2 terminal of the function switch to the 6 V terminal on the OX board. Connect another wire from the number 2 terminal on the function switch to terminal 2 on the AOS board. Connect a wire from terminal 3 on function switch to terminal 5 on the AOS board.

If a small six-volt battery or a holder with four pen cells is used for power, these can be mounted on the right rear corner of the lower chassis. Make a strip clamp, or bolt the holder directly to the chassis. This completes construction of the signal generator.

supplementary data

Specifications of the vhf signal generator, from the International Crystal Mfg. Co. specification sheet, are presented in **table 1.** The literature supplied with the OX kit shows several rf coupling circuits. I use a 7-inch length of insulated number 18 wire in a phono plug for a vertical whip antenna. Good marker signals are received through the station antenna with this arrangement. You can also use a loop on the rf output to calibrate a grid-dip oscillator at the oscillator crystal frequency.

Interesting articles on the OX oscillator and EX crystals appear in reference.

references

1. Don Nelson, WB2EGZ, "A Look at the EX Crystal and Its Oscillator," *ham radio*, April, 1968, p. 60.
2. Howard S. Pyle, W7OE, "Mini-spotter Frequency Checker," *ham radio*, May, 1969, p. 48.

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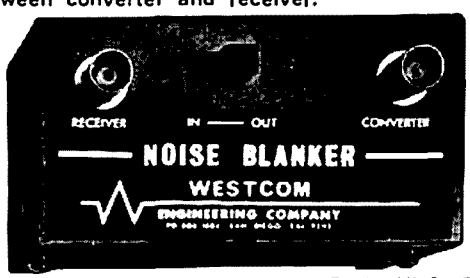
TNB-250 shown installed on a Swan 250 at accessory socket location.

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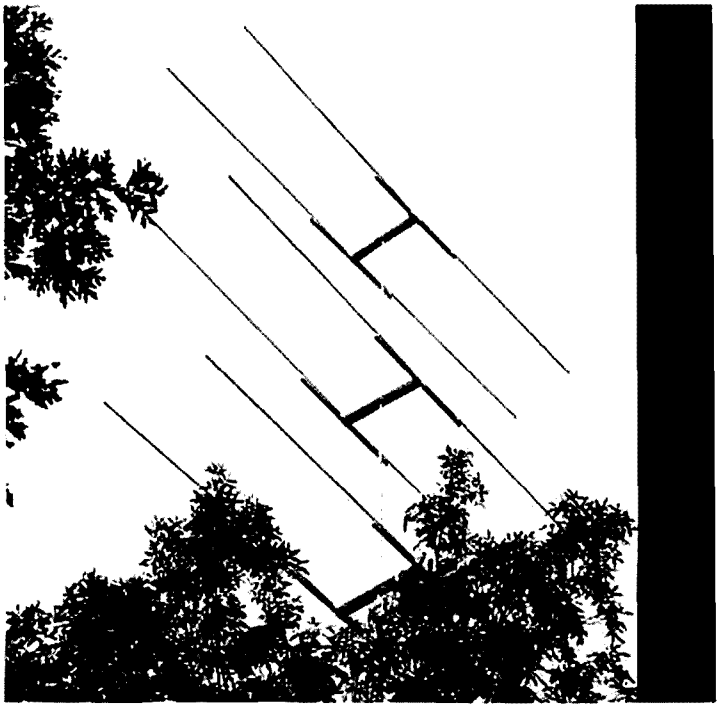
Refer to the New Products column of the August '68 issue of Ham Radio Magazine for additional information on the TNB Noise Blanker or write for technical brochure.

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big beam for six meters

Practical construction details of a 12-element collinear array for high performance on 50 MHz

John Stanley, K4ERO, 1501 Robin Road, Maryville, Tennessee

With more and more people getting on vhf in a serious way, vhf antennas have become a lively topic of conversation. One antenna that has always been popular with the 2-meter-and-up enthusiasts is the collinear; this is not without reason.

The collinear doesn't have quite as much gain, element for element, as a Yagi but it has two important advantages: low-Q operation and broad radiation pattern. The low-Q feature means that you can move over large parts of a band without fear of high swr. The broad radiation pattern means that aiming is not as critical; with high gain arrays this is a real advantage.

Although collinears are very popular on 144, 220 and 432 MHz, very few seem to be used on six meters. This is unfortunate because a rotatable eight-, twelve- or sixteen-element collinear for six meters offers one more advantage; it's easy to assemble up on

top of the tower. Since no part of the antenna has to be more than an arm's length from the tower itself, you don't have to balance a 20-foot boom or try to wrestle 50 pounds of aluminum tubing up through a maze of guy wires.

six-meter collinear

I hit upon the idea of a collinear for six when I was contemplating what kind of antenna I could put on my own 50-foot tower —a handy hickory tree. Obviously a long Yagi was out of the question. If I built it on the ground, I would never get it up through the branches, and I wasn't about to climb hand-over-hand out on the boom to fasten on the elements while floating between heaven and earth.

The 12-element collinear beam uses a handy hickory tree for support.



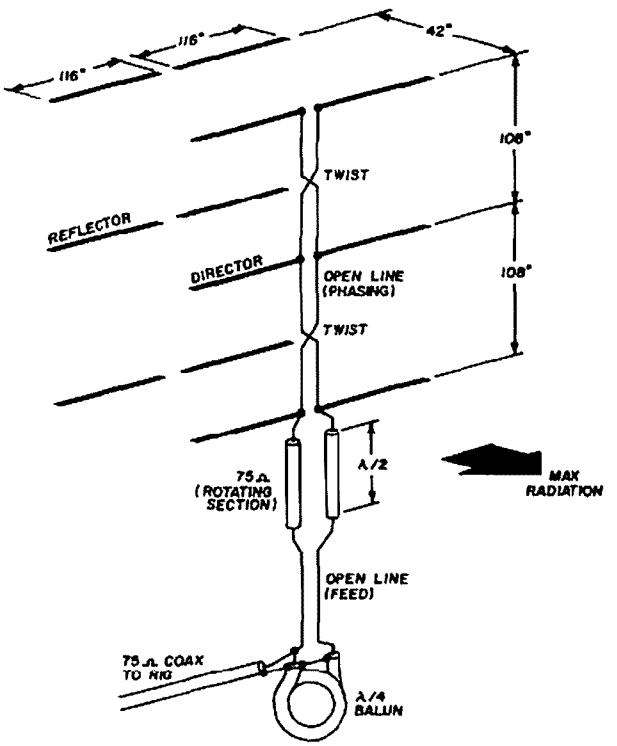


fig. 1. Electrical construction of the 12-element collinear for 6 meters.

With a collinear I could build each bay at tree top level without leaving the security of the hickory trunk, then raise each bay in turn by running the pipe mast up nine feet, and build the next bay. It was a snap. With a minimum of danger, work and investment, I had the most exciting six-meter beam I have ever used.

I am not going to give you all the mechanical details because the average ham should be able to equal my design or improve upon it.

construction

The booms are made of redwood, the elements are electrical conduit, and the insulators are ceramic rods and standoffs I picked up locally. Plexiglass is an excellent substitute for the insulators, aluminum would result in much lighter (and more expensive) elements, and oak might make more rigid

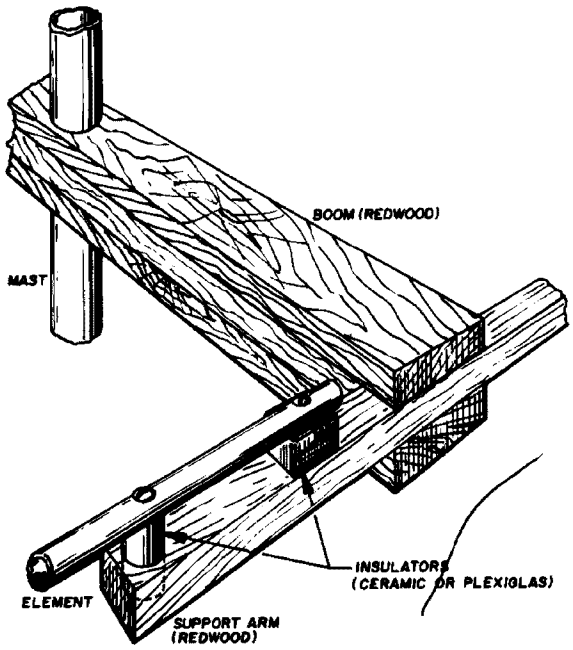


fig. 2. Mechanical construction used by K4ERO uses electrical-conduit elements, surplus insulators and redwood booms and support arms.

booms. I built my beam for under twenty dollars, and it has been up for more than a year.

Each driven element is 110 inches long and each reflector, 116 inches. The booms are 3½ feet long and stacked 9 feet apart (see **fig. 1**). The balanced phasing sections and balanced feedline are transformed to 75 ohms unbalanced through a balun. The rotating section consists of two half-wave sections of coax spliced into the open-wire line in the vicinity of the rotator.

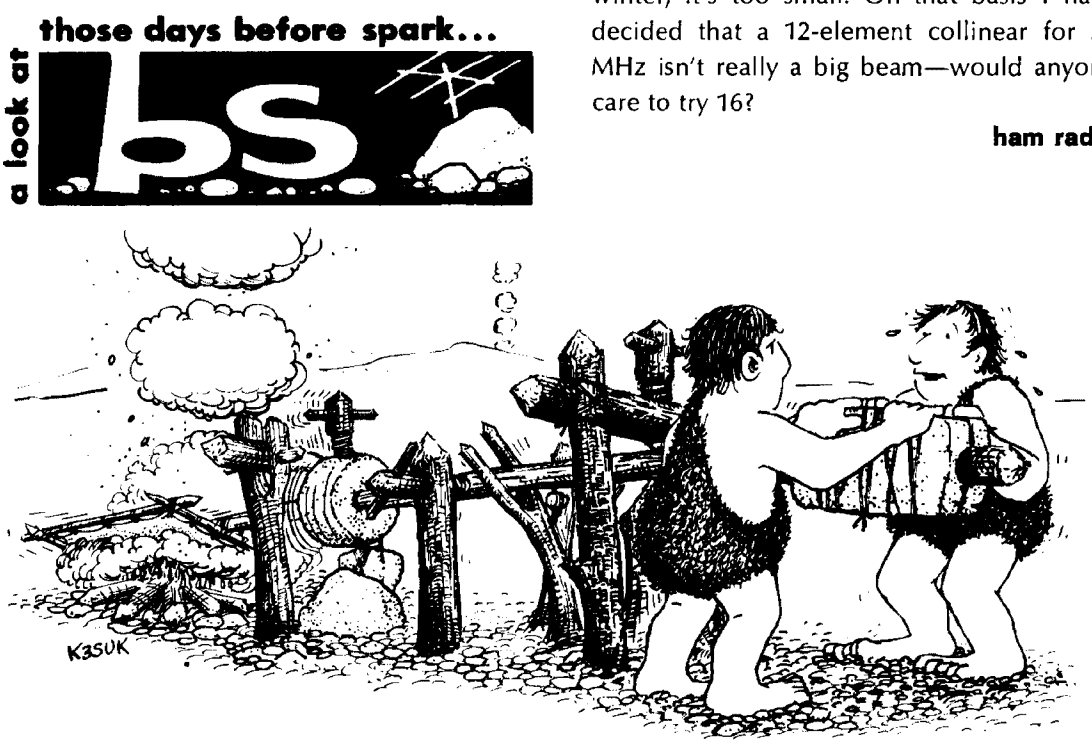
results

The performance of this array has more than exceeded expectations. The beamwidth is about 60° at the 6-dB points, and the front-to-back ratio appears to be near 30 dB. The swr is less than 2:1 over the **entire** six-meter band; from 50 to 51 MHz, swr is negligible.

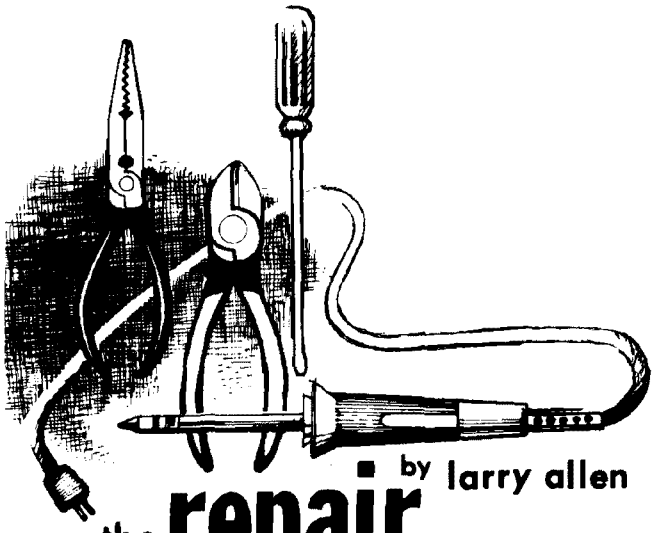
If there's a signal there, this antenna will capture it. I consistently work stations 300 to 400 miles away and can work over 200 miles anytime with 15 watts of single sideband. I have worked K4GXM (20 miles away) when he was running 1 mW of ssb to a 3-element beam. Scatter signals are very much in evidence on the low end of the band almost anytime, and if the band opens up suddenly, the wide beamwidth of this array means you are more likely to hear signals without careful aiming.

They say if your vhf beams stays up all winter, it's too small. On that basis I have decided that a 12-element collinear for 50 MHz isn't really a big beam—would anyone care to try 16?

ham radio



"How about you taking the "dah" side for awhile?"



by larry allen

the repair bench

tuning up ssb transmitters

If you missed last month's repair bench I recommend you go back right now and read it. If you've forgotten what it said, review it briefly.

For the working methods I outline in the next few pages, I assume you know how the equipment works. The explanations are only detailed enough for clarity. If you don't understand ssb, you shouldn't try a full alignment job anyway. Likewise, manufacturers' instructions seem like Greek unless you know how ssb works. (Editor's note: An excellent explanation of single-sideband appears in a series of ssb articles beginning with March, 1968 **ham radio** and running through February, 1969.)

getting ready

Here are a few items you want to make sure are handy on your repair bench before you start an alignment job. You'll need a vtvm with an rf probe, rf signal generator, separate audio generator, oscilloscope, and alignment tools to fit the adjustment screws and transformer slugs.

Gather the equipment around you, on a cart or on the shelf of the bench. Get it turned on, warmed up, and checked out to make sure it works okay.

Warm up the transmitter, too. Heat up all the tubes, and key it briefly into the dummy load a few times. However, if you're aligning it from scratch, don't key it until certain things are checked.

My example is a triple-conversion transceiver—you don't see too many of them, but they're great for demonstration. Many of its stages and arrangements are typical of other brands and models. There's a block diagram of it in **fig. 1.** It's an SBE model SB-34. Many SB-34 stages do double duty—more than in most transceivers. For the most part, I'll be concentrating on the transmitter.

However, this double service brings up something important: align the receiver first, whenever you have a transceiver on the bench. Transmit operation may depend on proper alignment of receiver stages. If you always align the receiver before you go ahead with transmitter alignment, you'll save yourself some going back to do part of the job over.

Aligning the SB-34 receiver is the same as any triple-superhet. Just be sure you do it accurately. Later, transmitter alignment depends on how well you aligned certain mutual stages.

Presetting the operating (front panel) controls is an important first step. The manufacturer's booklet usually lists them. For the SB-34, be sure **pa tune** and **pa load** are fully counterclockwise (ccw). Turn **mike gain** R1 all the way down. The sideband switch goes at **usb.** The front-panel meter should be switched to read plate current of the final, the stage most likely to be damaged if tuning adjustments aren't carried out properly.

Internal adjustments to preset in the SB-34 are voltage regulator (not shown), final bias pot R81, neutralizing capacitor C86, and such. As I said, the manufacturer's alignment instructions list the preset controls and tell how to set them.

Certain voltage tests should often be made before the transmitter is keyed on for tuneup. Check plate and screen voltages on the finals and driver, the bias source for the finals, and any regulated voltages (the vfo B-plus supply is regulated in the SB-34).

Measure normal B-plus, with the unit on standby or in receive mode.

getting to work

Now for some alignment. Begin with the bandswitch set for the highest band, and the vfo at the high end of its dial. With no mike input and **mike gain** full ccw, adjust the bias two outputs as shown in **fig. 2.** Build the simple three-resistor isolation network.

Identify the frequency of the audio tone from your rf generator. Set the dial of the audio generator to furnish a signal about 1 kHz higher in frequency. Adjust the output controls of the two generators to supply the same level of signal.

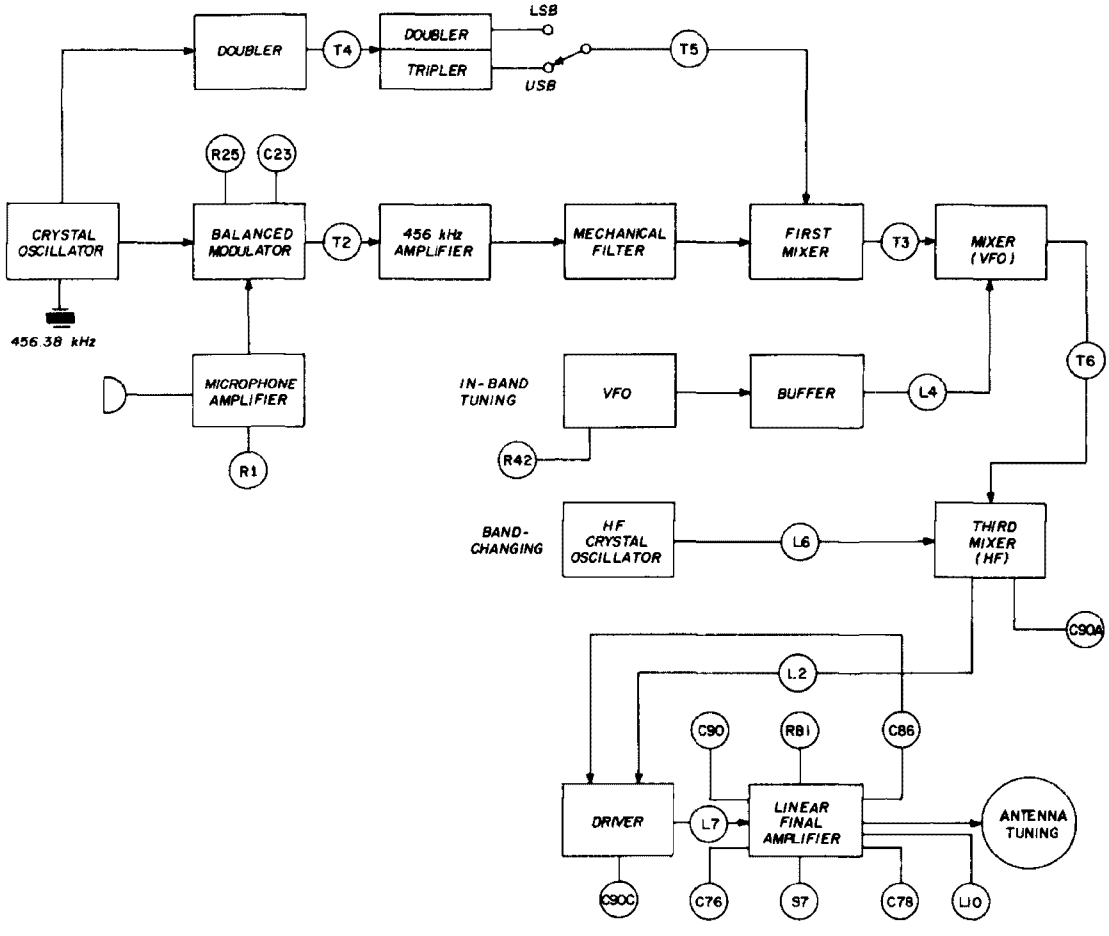


fig. 1. Abridged block diagram of SB-34 ssb transceiver. It is used as an example of how alignment is done.

pot for the finals so the plate-current meter on the front panel reads about 1.2 mA. Plate current in the finals with no modulation should be no more than 1.5 mA and no less than 1.0 mA.

Next job is to adjust the **pa tune** and the **pa load** knobs, which are operating (front panel) controls on the SB-34. This is done best with a two-tone signal going through the transmitter.

Use the audio generator and the audio signal from your rf signal generator. Mix the

Feed the two signals into the mike jack of the ssb transmitter. In the balanced modulator, they generate two sideband signals about 1 kHz apart. If the two tones are pure sine waves, with no significant harmonics, they produce in the transmitter two **rf** signals near the output frequency but 1 kHz apart.

An oscilloscope connected to the output of the transmitter displays the pattern in **fig. 3A.** The scope's sweep should be set to about 500 Hz, to lock in two full cycles of the beat note (which is somewhere near

1000 Hz). A scope connected across the dummy load is handy to watch for signs of instability during adjustment.

adjusting the final

With the two tones applied, turn **mike gain** up just enough to make a slight reading on the rf output meter. Adjust **pa tune** for a dip in the dc plate current meter on the front panel. Then, watching the rf output meter, adjust L7 and C90C for maximum rf output; they are driver tuning adjustments. Again dip the final plate current with **pa tune.**

A little at a time, turn **pa load** clockwise to increase rf output. After each increase, turn **pa tune** slightly counterclockwise and then clockwise till the deepest dip is obtained on the dc plate current meter.

As you advance the loading, watch the scope for signs of instability. You can also see them in erratic actions of the meters. Instability means you need to adjust neutralizing trimmer C86, which feeds back some rf signal to the driver.

It's even better to neutralize as you go along. After every couple of advances in loading, after you've re-dipped **pa tune,** touch up C86 so the dip on the dc plate meter coincides with the rf output peak on the output meter. That keeps the stage neutralized properly and prevents it breaking into oscillation at all.

Load the final until there is no further increase in rf output when you turn the **pa load** knob further clockwise. Then turn it back ccw just enough to reduce rf output noticeably. Re-dip the **pa tune** knob, check neutralization, and the linear final is tuned.

Caution: Never keep the transmitter keyed on more than a few seconds with the two tones applied. It's bad for the driver and final tubes.

Go back to the output of the last mixer. Adjust L2 and C90A for maximum rf output. Do the same again with L7 and C90C. (C90B —not shown—was adjusted during receiver alignment.)

for linear operation

Now is a good time to check linearity of the final, mixer, and driver stages. Turn up the mike gain until the sine waves on the scope begin flattening out on top like those in **fig. 3B.** With mike gain not quite making flat tops, recheck **pa tune,** neutralizing, and mixer and driver adjustments. Also make sure the **pa load** knob has a slight leeway to be

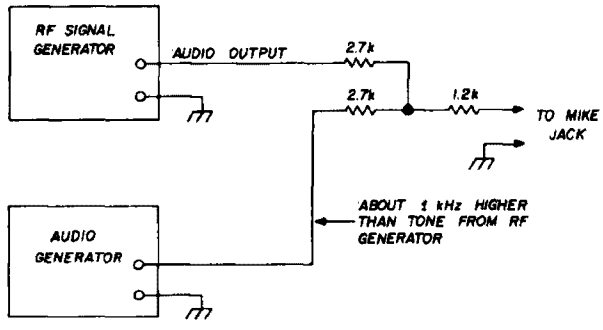


fig. 2. Build the mixing pad to keep the two generators isolated. Set their outputs for the same level; scope is a good indicator.

turned clockwise and still cause an increase in rf output. However, always leave it slightly ccw from maximum rf output.

Rf output from the SB-34 should be at least 50 watts (about 50 volts across a 50-ohm load) in this test. If, to get that much output, you have to advance **mike gain** so much that the sine waves are flattened out, there is nonlinearity in the transmitter. If not, go on to the voice test.

Disconnect the two-tone mixer resistors and plug the mike into its jack. At a normal voice level, say "a-a-a-h-h-h" into the mike. Turn up the mike gain until flat-topping is obvious in this new scope pattern. Back **mike gain** down enough to cure the flat tops and mark the position. That's the point of most linear operation. Rf output should be about 50 watts again.

Now turn the mike gain down completely. Turn **pa tune** clockwise to a peak on the output vtvm. Adjust 24-MHz trap coil L10 for minimum output reading. Turn up **mike gain** and re-dip **pa tune.**

balanced modulator

If you had any trouble with flat-topping before the rf output reached about half the unit's PEP output rating, something is amiss

in the transmitter. Most likely trouble spots are the balanced modulator and the linear final. Check the modulator first. The stage is shown in **fig. 4.**

Turn **mike gain** fully ccw. You want absolutely no modulation. First adjust the potentiometer for the least *rf output* across the dummy load. Then do the same with the capacitor. Both are labeled **carrier balance** in the diagram. In some transmitters one may be called **carrier null.**

Flip the **usb/lsb** switch back and forth. The rf output reading should be the same in either position, and very low—in the SB-34, only a fraction of a volt across the 50-ohm dummy load. If the reading changes as you flip the switch, readjust the slug in T2, the transformer at the output of the balanced modulator. Then again set the pot and trimmer for least rf. Find the T2 setting that leaves the sidebands balanced when rf is minimum.

checking other bands

The SB-34 is a multiband transceiver covering 15 through 80 meters. You'll need to align the other bands.

Set the bandswitch to 40 meters and tune to 7.075 MHz. Advance **mike gain** to mid-range. Adjust the 40-meter (rear) slug of L7 for maximum rf output. Tune up **pa tune** and **pa load** as usual. You shouldn't need to change the neutralizing trimmer.

Switch to 80 meters and tune for 3.8 MHz. The same slug in L7 affects 80 meters. Adjust it to split the difference between the output reading you got on 40 meters and the one you get on 80. With **mike gain** up almost to flat-topping, adjust **pa tune** and **pa load.** Output should be about 50 volts across the 50-ohm load.

Switch to 20 meters and tune to 14.25 MHz. Advance **mike gain** and tune up **pa tune** and **pa load.** At flat-topping, output should be 50 volts or more.

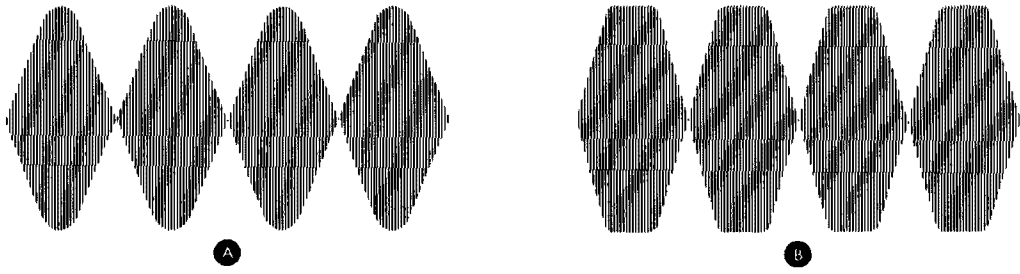


fig. 3. Four cycles of two-tone beat signal in rf output of ssb transmitter. A shows normal full modulation; B shows flat-topping.

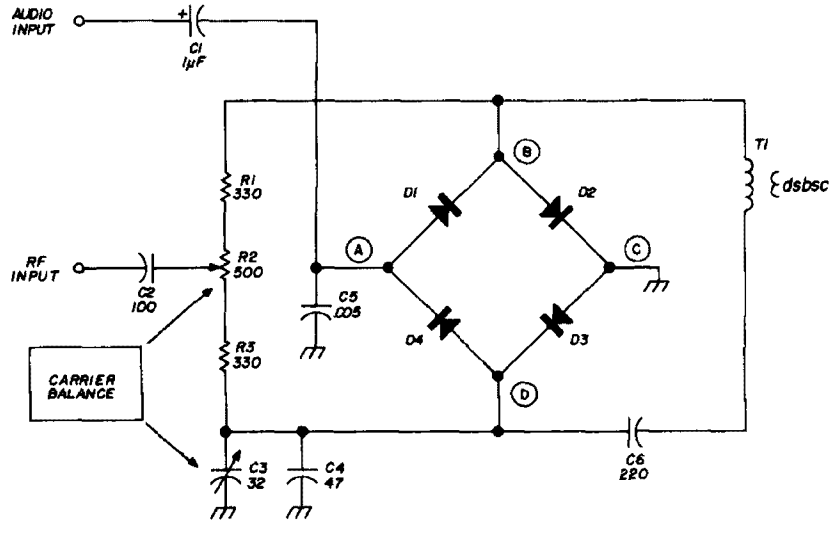


fig. 4. Balanced modulator in SB-34 is very typical. Balancing controls may have different labels; some models may have two trimmers.

In most ssb transmitters, only one oscillator is variable; the others are crystal controlled. In **fig. 1,** you can see several adjustments in those crystal oscillator and multiplier stages. You need only one simple rule of thumb to align them.

Connect an rf probe and vtvm to the driver stage. Be sure receiver alignment has been done. Work backward, aligning in this order: L6, T6, L4, T3, T5, and T4. Adjust each one for maximum rf measured at the driver. These adjustments are merely peaking devices for their respective signals; they're adjusted to pass along the most rf to the main transmitter stages.

A few transmitters have trimmers in the oscillators, to warp crystal frequencies precisely. They can be adjusted without a frequency meter, **if** the rest of the alignment has been precise. The maximum-rf system works for them, too.

An exception is the variable frequency oscillator, which determines output frequency. Fortunately, frequency precision isn't serious in the ham bands unless you're working near the edge.

Many vfo's have trimmers to calibrate the dial. This is done best in the **cw** or **tune** mode of the transmitter, so a carrier is being generated. The sideband could cause confusion. Best test indicator is a frequency meter, or a receiver that has been calibrated against WWV recently.

The way most transceivers develop their output frequencies—by a form of synthesis—the vfo has only one band to be calibrated. Then proper tracking among the various bands becomes a matter of making sure the crystals in the band-changing oscillator (see **fig. 1**) are accurate. Again, a frequency meter or calibrated receiver is the best indicator. Some transceivers have a 100-kHz calibration oscillator; you calibrate the receiver mode with that, and then check the transmitter against it.

other transmitters

Most of these principles outlined apply to all ssb transmitters. But of course there are exceptions, when transceivers have unusual stages. Study the schematic and block diagram of any transmitter you have to align. If the instruction manual doesn't have a block diagram, draw one; it'll help you understand what you're doing during alignment.

One typical exception is the 9-MHz carrier oscillator adjustment in a Gonset transceiver. If it's wrong, the sidebands are offset and distorted because they miss the bandpass curve of the filter that follows. There's an abbreviated block diagram of the arrangement in **fig. 5.**

If the oscillator puts the sideband off to one side of the filter's response curve, the amount of signal reaching the first transmitter mixer is less than normal. So the rf voltage at the emitter of that mixer is a good indicator. You adjust C23, the carrier oscillator trimmer, until rf at that point is exactly 65 mV (0.065 volt). More or less means the carrier oscillator is off frequency.

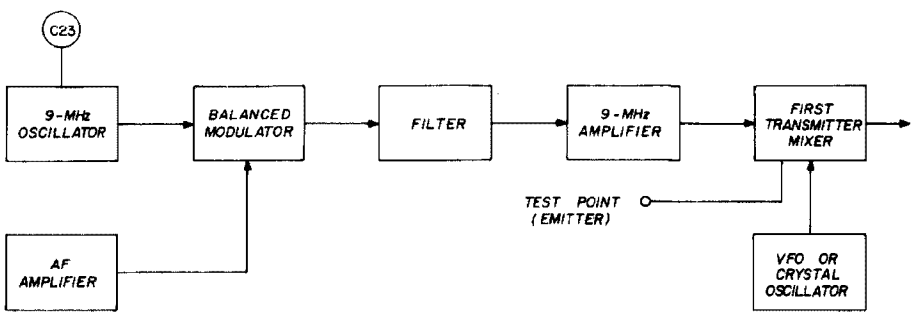


fig. 5. Early stages in transmitter using 9-MHz carrier frequency. In Gonset model, frequency trimmer (C23) is set according to rf voltages at test point after the filter.

linear amplifiers

The steps you followed with the final stage of the SB-34 are similar to those for any high-power linear final. However, there are precautions.

Set the bias before you do anything else. You may refine its adjustment later, but the main thing is to keep the bias high until everything else is properly tuned.

Next thing is to get the plate tuning dipped as soon as you can. That keeps current low and protects the tube (or keeps the overcurrent relay from kicking out). Make sure the exciter is furnishing enough signal. You can usually tell with an rf meter across the linear-amp input. Remember that you can't judge drive from the grid voltage; a linear amp doesn't develop grid current the way a class-C amplifier does.

When you're loading a linear amplifier into the antenna, do it gradually. Watch the dc plate current meter as well as the rf power output indicator. Many linear amplifiers can be loaded well beyond their proper (or legal) operating limits.

After each load increase, re-dip the plate tuning. Notice whether the rf output peak and the plate-current dip occur at the same setting of the plate tuning capacitor. If not, neutralize the stage before proceeding. An oscillating 1-kW linear can make a lot of trouble both for the tube and for parts in the circuit. It can start arcs you'll find nigh impossible to stop.

next month

So much for tuning up and aligning ssb transmitters. With what you've learned here, you should have no trouble following manufacturers' instructions or working out an alignment procedure for your own homebrew ssb rig.

The subject of next month's **repair bench** applies equally to ssb, am and cw rigs. It's oscillators. True, ssb rigs have a lot of them, but no transmitter can operate without at least one.

When an oscillator quits, it can be one of the most difficult troubles to pin down. So, be here next time and I'll explain in detail what makes oscillators quit. **ham radio**



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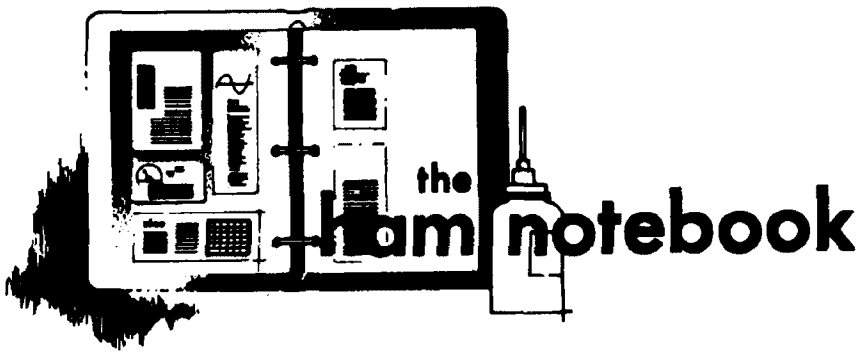
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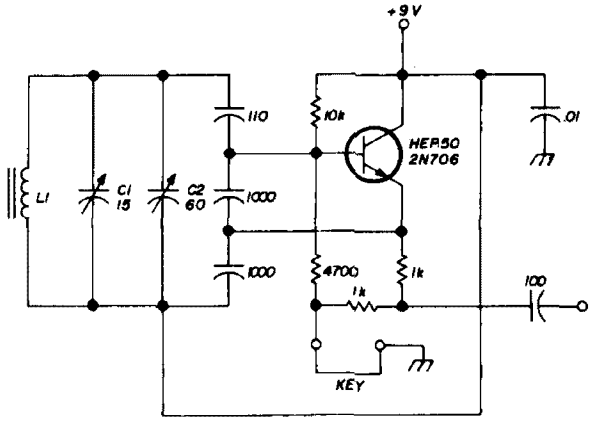


oscillators

The 40-meter Seiler oscillator shown in **fig. 1** was built by K1BBU as part of a low-power solid-state transmitter. Inductor L1 consists of 23 turns number 20 enamelled wound on an Amidon T68-2 toroid form (L = 3.2 μH). C1 is a 15-pF variable with a minimum capacitance of 2.3 pF (E. F. Johnson 148-1). C2, the 60-pF padder, is used for setting the frequency range. With these tuned-circuit constants the vfo tunes from 6.93 to 7.24 MHz. K1BBU reports that this circuit exhibits excellent stability and keying characteristics.

The Seiler oscillator circuit shown in **fig. 2** was built by W1DTY to see if the stability of a completely non-temperature-compensated Seiler oscillator was stable enough for 455-kHz ssb generation. Total drift in this circuit, as measured with a frequency counter, was 40 Hz, including warm-up drift. With a 33-pF N750 capacitor in parallel with the inductor, drift was less than 10 Hz from turn on. A 455 kHz crystal oscillator was turned on at the same time for comparison purposes; total drift of this circuit (shown in **fig. 3**) was 4 Hz.

fig. 1. Seiler oscillator tunes 6.9 to 7.2 MHz.



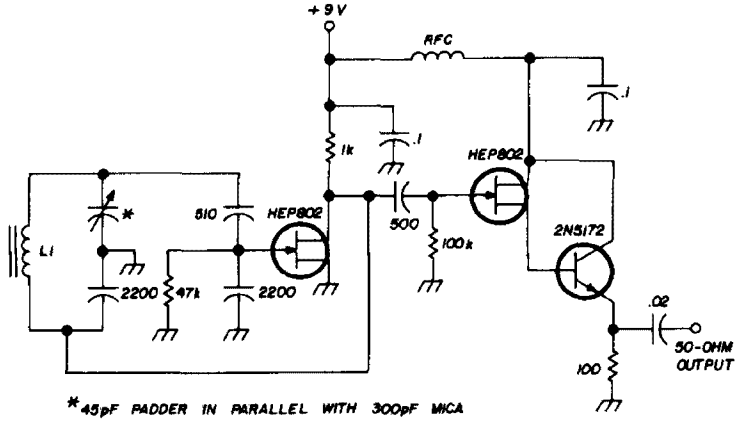


fig. 2. This 455-kHz Seiler oscillator exhibited less than 10 Hz drift when compensated with a 33-pF N750 capacitor.

Inductor L1 in the 455-kHz Seiler oscillator consists of 190 turns number 26 enamelled on an Amidon T68-2 form. When the 33-pF temperature compensating capacitor was added to the tank circuit, the 300 pF mica was reduced to 270 pF.

The buffer circuit in **fig. 2** provides excellent isolation for the oscillator and is capable of putting nearly 50 milliwatts of rf into a 50-ohm load.

The *simple crystal oscillator circuit* shown in **fig. 3** is an excellent circuit for crystals between about 70 kHz and 2 MHz. It presents a 32-pF load for the crystal. The voltage at the collector of the transistor should be approximately one-half

the supply voltage and can be adjusted by changing the value of the 100k base-bias resistor.

A simple crystal oscillator circuit for the frequency range between 2 and 20 MHz is shown in **fig. 4.** This oscillator is a modified Pierce circuit that provides good output and high stability. For proper operation the output of this circuit should work into 600 ohms or more. If the output load is less than 600 ohms, the fet-bipolar buffer circuit shown in **fig. 2** may be used.

overtone oscillator

The simple transistor overtone oscillator shown in **fig. 5** is designed for overtone crystals in the frequency range from 20 MHz to 100 MHz. The frequency of operation is determined by the tuned circuit. Capacitor C1 is a 25 pF trimmer with a minimum capacitance of 7 pF or less. Inductor L1 is chosen for the desired frequency as shown in the table in **fig. 5.** Most npn transistors with an f_T of 200 MHz or more will work in this circuit. In operation, the tuned circuit is adjusted to the overtone frequency of the crystal.

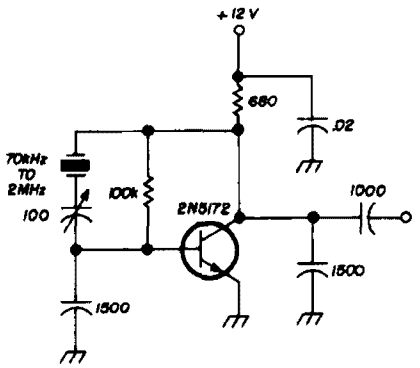


fig. 3. Simple crystal oscillator circuit for crystals from 70 kHz to 2 MHz.

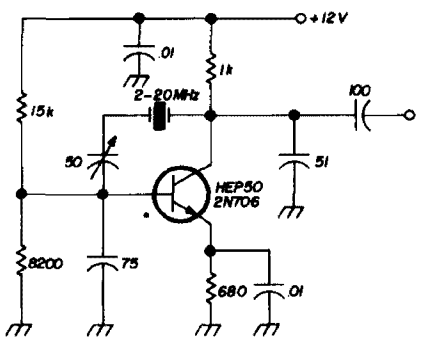
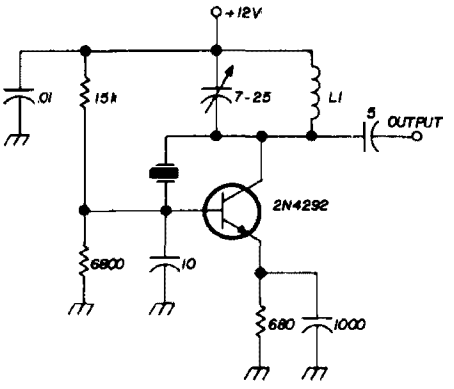


fig. 4. This modified Pierce oscillator is useful for crystals from 2 to 20 MHz. Output should not be loaded with less than 600 ohms.



frequency (MHz)	L1
20 – 35	2.4 μH (Miller 4606)
35 – 60	.68 μH (Miller 4590)
60 – 100	.22 μH (Miller 4584)

fig. 5. Simple overtone oscillator covers the range from 20 MHz to 100 MHz.

half-watt solid-state cw transmitter

I have a friend who ridicules solid-state equipment, even after I've worked both coasts from mid-U.S.A. on six-meter phone with 100 milliwatts! So I told him I'd build and give him a solid-state cw transmitter. Its schematic is shown in **fig. 6.**

I made it as simple as possible, using junkbox parts and surplus or bargain transistors. It has only one tuned circuit, the output. The complete circuit evolved from several other circuits plus some ideas of my own.

At first, some of the base loads were rf chokes, but due to feedback or chassis ground loops, the final "took off" when detuned, so resistors were tried (with the same dummy antenna load). The resistors remained, and the unit is very stable.

The rf chokes are peaking coils from an old tv set; transistors are from IBM or similar boards sold everywhere. The coil is from an old surplus receiver.

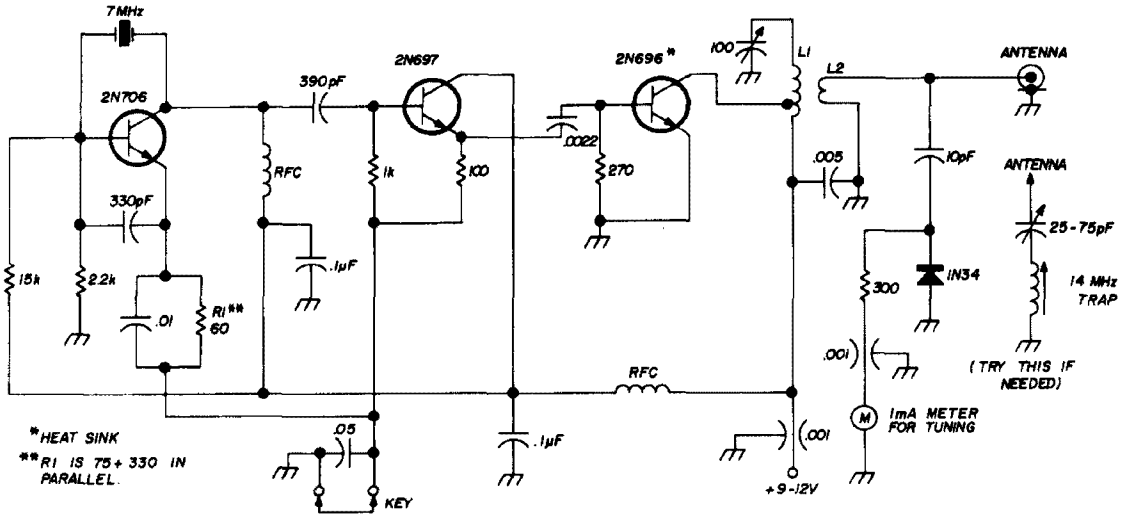


fig. 6. Solid-state cw transmitter built from junk-box parts. It puts out a half watt using a 12-volt supply.

The unit may be run at 15 or 16 volts, but all transistors should have heat sinks. It would be a good idea to change the 15 kilohm resistor in the oscillator base if you plan to run more than 12 volts. (Try 22 kilohms to 27 kilohms.)

The unit was connected to a Bird wattmeter and indicated 200 mW at 9 volts and 500 mW at 12 volts, with a total current drain of around 100 mA at 12 V. You'll probably get more output with a 2N697 in the final, because it has more gain than the 2N696. In my case, the 2N697 worked better in the driver stage. This will light a number 47 bulb to about normal brightness.

If the second harmonic is bad, try the trap shown. Dip it with a GDO, or insert a 15-MHz crystal in the unit and adjust the trap for no output, using your receiver. (The trap wasn't installed on his transmitter.)

Bill Esiick, KØVQY

three bands with two linears

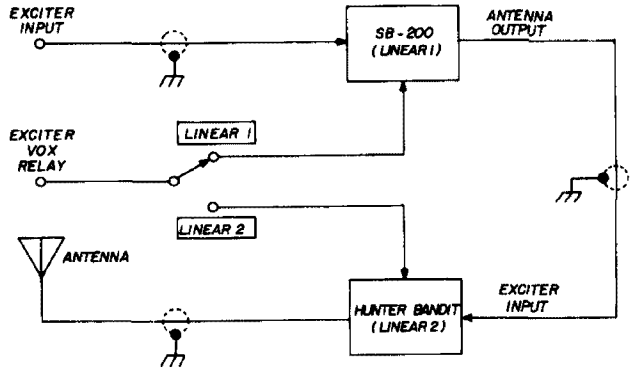
Having two linear amplifiers is a sign of affluence, but they're mighty handy when there's a need to change bands quickly. I had a Hunter Bandit, but found a Heathkit SB-200 at a very attractive price, so I bought it. It occurred to me that here was a way to have one final for 10 and 15 meters (the SB-200) while the Hunter was on 20.

However, a question immediately arose: How can you have both linears in the line without a batch of switches or relays, or without manually changing coax cables? A brain storm gave me the answer, although at first I had some trepidations about trying it.

Each linear has an internal antenna relay that permits the exciter output to go directly to the antenna if the power to the linear is not activated. With the linear power on, the linear output goes to the antenna, while the exciter drives the linear. As shown in **fig. 7,** I connected the exciter to the input of one linear and the output of the second linear to the antenna. Two ac outputs from each linear are always connected to the line.

Thus, if I turn on the power to linear **1,**

fig. 7. Arrangement for separate amplifiers to give dual-band coverage. A spdt switch switches exciter vox between amplifiers.



its output goes through to the antenna output of linear **2** and out on the air. If I put the power on linear **2**, the exciter feeds through the relay in linear **1** to the input of linear **2. Just don't put both linears on at once!**

The only other addition is a small switching arrangement to enable you to switch the output of the exciter vox-controlled relay to one or the other of the linears. As they run on different principles, I couldn't figure a way to avoid that one switch. However, someone may come up with an idea.

Gay E. Millius, Jr., W4NJF

low-voltage zener tester

With the advent of transistors and IC's, it's desirable to have low-voltage power supplies and transistor testers for use on the workbench. The tester described here is not the ultimate in zener testers, nor do I make any claim for its originality. But it does work, is very accurate, and the cost is within the reach of most serious hams and experimenters. It will measure zener voltage up to 10 volts at the desired current from 15 mA to 40 mA.

I must point out that you can't see the response curve of the device (a curve tracer is needed for this); but if you're certain the device is a zener, this tester will measure its breakdown voltage and allow selection of precise voltages. You can also check voltages of unmarked and surplus devices.

For testing low-voltage zeners, I feel it's desirable to use a constant-current source. A standard parameter used by many companies for 400 mW, 2.2- to 10-V zeners is 20 mA.

The voltmeter consists of the 100-μA meter 1 and its 100k multiplier. The accuracy of your tester will be determined by the accuracy of these two components.

I used a 100 μA API meter with 0.5 percent accuracy. Less expensive parts can be used if you're not too concerned with accuracy. The meter should be no larger than 100 μA, though because its loads the circuit.

To set up the tester, merely plug it in and short circuit point A to point B through a 50-mA meter. (**Yes** short circuit—this is a constant current source.) Adjust the 200-ohm pot for the desired test current. (I use 20 mA for 400 mA zeners.) Don't be concerned when the meter reads off scale without a zener across terminals A and B. The maximum output to the meter is only about 12 V dc, which means it's only about 20 percent overscale. This won't damage the meter.

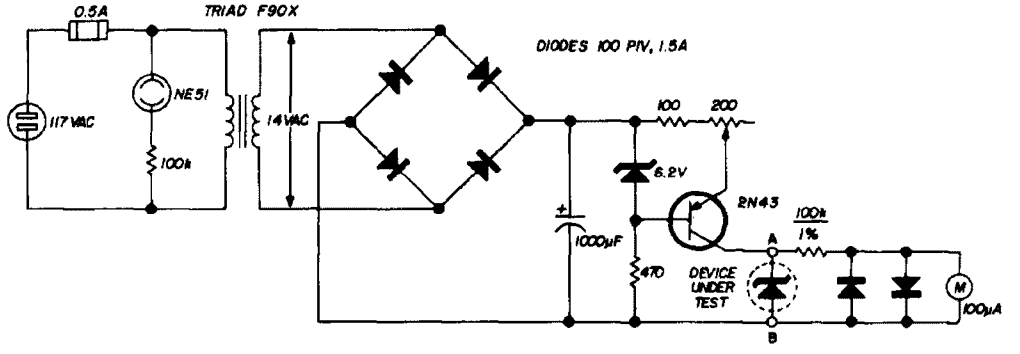
After this adjustment, place the zener to be tested across test points A and B. The voltmeter will drop to the zener voltage at the set current.

If a unit should be placed backwards across points A and B, the voltmeter will drop to about 0.8 volt.

The layout of components isn't critical. I built my tester on vectorboard and placed it in a Bud 4x5x6-inch minibox.

M. Weinschenker, K3DPJ

fig. 8. Zener-diode tester uses constant-current source that is adjustable from about 15 mA to 40 mA. Zener voltage is read on the voltmeter consisting of the 100-microampere meter and 100k multiplier.



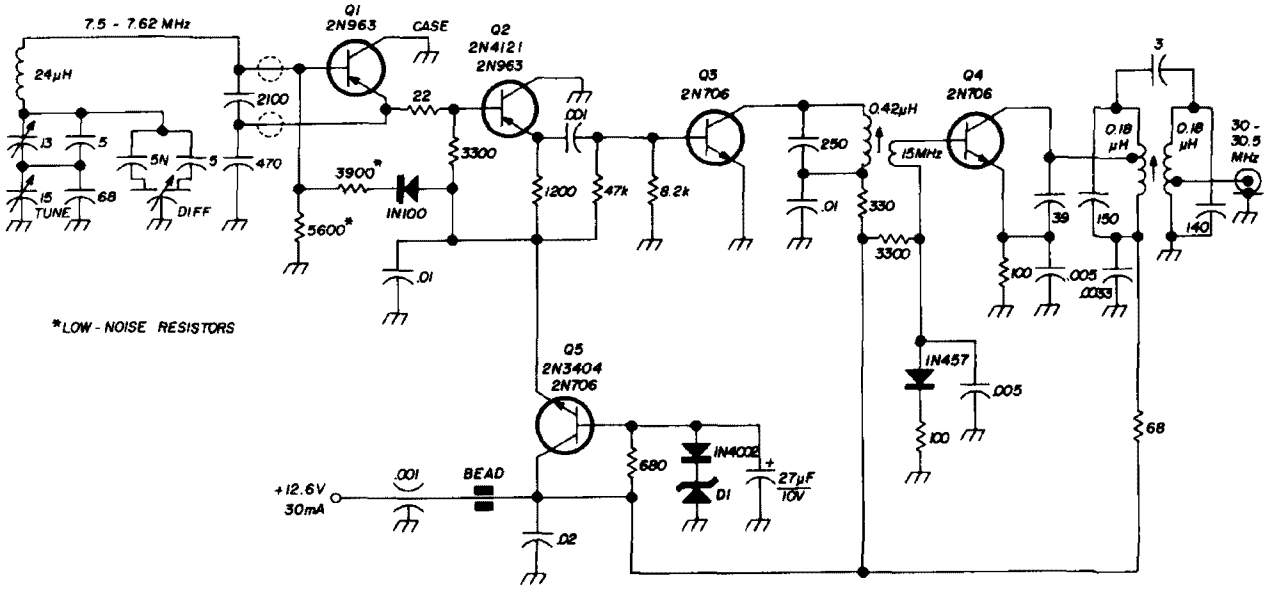


fig. 9. Vfo used by W1OOP as part of his two-meter system. Diode D1 is an 8.4 to 9.3 volt reference diode such as the 1N935A or 1N3154; temperature coefficient is important. According to calculations, the power dissipated in the oscillator tuned circuit is less than 3 mW.

vfo transistors

When choosing a transistor for a vfo, several characteristics should be considered: power gain, thermal resistance and capacitance. Transistors with high power gain allow loose coupling to the resonant circuit so transistor changes have less effect on frequency; high f_T ratings imply high power gain.

Thermal resistance characteristics (degrees rise per milliwatt) are important because transistor characteristics and capacitances drift with temperature; less

fig. 1. Oscillator transistors listed by figure of merit.

JEDEC Number	Type		f_T (MHz)	Pin (mW)	C_{ob} (pF)	Thermal Resistance °/W	Figure of Merit	Remarks
2N1141	pnp	ge	300	5	2	100	400	class C
plastic 2N918	npn	si	400	10	1.7	200	154	slow heating
2N963	npn	si	300	5	2.5	250	150	low-cost switch
2N700A	pnp	ge	400	5	1.4	600	110*	vhf
2N2368 40405	npn	si	300	10	3.5	150**	79	saturated switch
2N4121	pnp	si	400	10	4.5	250	76	slow drift
2N3866	npn	si	600	50	4	34***	60	class C
2N918	npn	si	400	10	1.7	580	53	vhf
2N706 40404	npn	si	200	10	5	150**	45	saturated switch, some noisy
2N1742	pnp	ge	500	12	2.5	1000	27	slow heating
2N3250	pnp	si	300	50	6	150	16	general purpose

* When biased at 5 V, 1 mA; figure of merit drops to 70 when device is biased to 5 V, 2 mA.

** Use thermal resistance of 200°/W because of polarity.

*** Use thermal resistance of 100°/W because of polarity.

temperature rise means less drift. Lower thermal resistances are found when the collector is tied to the case, so this is often an important consideration.

Some applications require appreciable power output from the vfo, but if you don't need it, low-power operation is preferred. Pick a transistor that has high gain at low voltage and current; some types such as "forward-agc i-f and rf amplifiers" are poor performers at low voltage. Most transistors that are characterized for class-C operation are good. Computer types are usually good, but silicon "saturated-switch" types may be noisy at low frequencies. Power level for crystal-controlled oscillators is limited by crystal heating; high f_T minimizes this.

Other important considerations in vfo design are transistor size and polarity and supply voltage. Smaller transistor packages are preferred; as far as polarity goes, pnp transistors are preferred for grounded-collector operation in mobile power supplies in American cars. Mobile operation is also an important consideration when discussing power supplies: automotive battery voltage means a maximum of nine volts regulated—well regulated—to the oscillator.

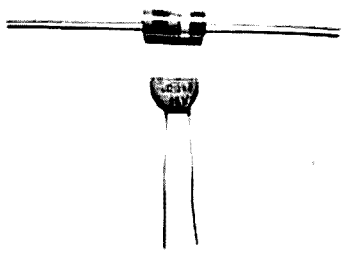
The best way of choosing a semiconductor is to establish a figure of merit for oscillator transistors. This figure of merit should be proportional to f_T, and inverse as the square root of the output capacitance at the operating voltage, inverse as the thermal resistance and inverse as the dc power input at the operating point. The transistors in **table 1** are listed according to this figure of merit.

The vfo transistor in the circuit of **fig. 9** is a 2N963, an inexpensive switching transistor with an extremely good oscillator figure of merit. In the circuit I built, emitter voltage on the 2N963 is 5.3 volts and emitter current is 1.1 mA; total input power is 5.8 mW. Since the thermal resistance (junction to case) is 0.25°C/mW, the temperature rise of the semiconductor junction in this circuit is 1.46°C maximum. Power dissipated in the tuned circuit is less than 3 mW.

The basic 7.2 to 7.5 MHz output of the vfo drives the doubler stage Q3; the output from Q2 drives another 2N706 doubler, Q4, to 30 MHz. The 30 to 30.5 MHz output is used with a 114-MHz crystal controlled source in a conversion scheme to provide vfo control on two meters.

Hank Cross, W1OOP

trimmers



The components illustrated above have not been ruined. This simple technique of adjustment may be old hat to electronics technicians, but it always seems to surprise hams. It is especially useful when breadboarding with junk-box components.

Resistors may be adjusted upward in resistance as much as 50% with a file, as shown; the cut illustrated raised this resistor by about 20%. With this method it is possible to make your own 1% resistors or to match a set of components. The filing can be done with the resistor connected in a bridge or other active circuit; remember to start with a resistance lower than the desired value. For reasonable changes, the reduction in power-handling capacity is small.

Ceramic capacitors may also be trimmed in the same manner, with a file or grinder. Practical adjustment range is 50%, and the illustrated capacitor measured about .0025 pF. In-circuit changes are possible.

After reaching the desired value, brush or wash away any dust, and apply a coating of insulating varnish or spray. With careful preparation, these "precision" components are as good as the expensive kind.

Dale E. Coy, W5LHG

CQ de W2KUW

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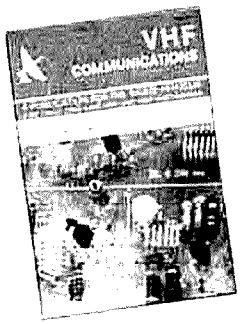
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3CX2500
3CX3000
3CX5000
3CX10,000
3E29
3K (any digits)
4-65
4-125/4D21
4-250/5D22
4-400A/8438
4-1000A/8166
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4CX250R/7580W
4CX300
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4CX10,000/8171
4X150G/8172
4PR60A or B
4PR (any digits)
5-125B/4E27A
5R4WGB
6BL6
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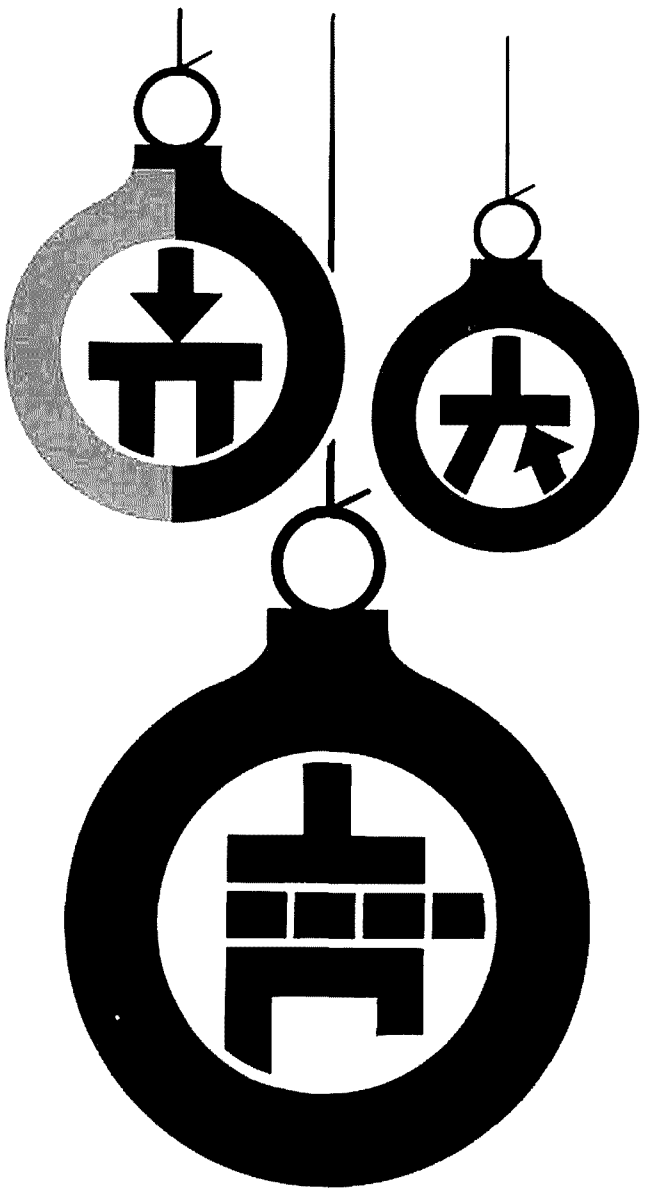
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staff

editor
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vhf editor
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fm editor
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associate editors
A. Norman Into, Jr., W1CCZ
Alfred Wilson, W6NIF
James A. Harvey, WA6IAK

art director
Jean Frey

publisher
T. H. Tenney, Jr., W1NLB

offices

Greenville, New Hampshire 03048

Telephone: 603-878-1441

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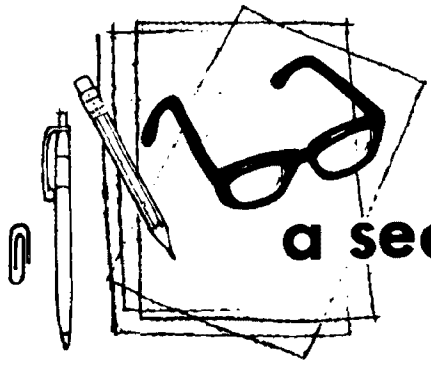
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a second look

by jim fisk

The editor of a technically oriented magazine like **ham radio** wears several hats. I could occupy the whole page describing the details that require attention to keep the magazine running smoothly. But I'd like to talk for a moment about one very important editorial task that means the difference between a technical magazine that instills confidence in the reader, and one that doesn't. This task, which I share with my staff, is that of researcher and seeker of truth.

Most of the articles published in **ham radio** are contributed. Authors range from enthusiastic hams who want to share an idea, to fellows with engineering backgrounds (who also want to share an idea). I think this is fine, and I welcome the output of everyone who is interested in contributing something that will benefit all hams.

Budding authors often ask, "What kind of articles are you looking for?"

That question is difficult to answer since many new manuscripts come in every day, but generally speaking, I am looking for simple solid-state construction projects that the average fellow can build in one or two weekends. Larger construction projects are also welcome, but the average **ham radio** reader must split his leisure time between amateur radio and other interests, so he doesn't have the time to build a Chinese copy of a complex piece of equipment.

When I read an article contributed to **ham radio,** the first thing I look for is interest value. If the contribution passes this test, the next thing I look for is technical accuracy and attention to detail.

The contributed article doesn't have to be a literary masterpiece. If you have a good idea; if it's well documented; if the illustrations and technical discussion are clear and accurate—you probably have a winner!

My staff and I are fairly adept at ferreting out technical oversights and inconsistencies, but of course, we can't catch them all. Despite research and keeping the mailman busy between our editorial office and the author, errors occasionally do creep into the magazine. As a contributing author, you can help reduce the number of "short circuit" items that are published from time-to-time.

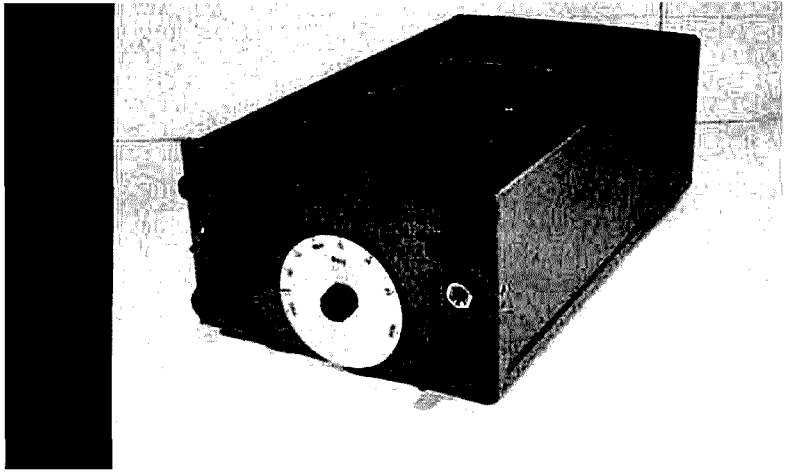
For example, when you draw your illustrations, go over each component, each wire, and each connection. Take another look at numerical values. A misplaced decimal point in a component value can mean the difference between a circuit that works and one that can cost the builder a small fortune in burned-out parts.

Not only does this cause considerable misery to the builder, but as publishers **we** are taken to task for the error. Result: a "short circuit" item must be published in a later issue. This takes care of the initial problem, but by then, considerable damage could have been done.

If you prefer to read **ham radio,** rather than write for it, you can help by telling me the kind of article that you like. If you have a pet project in mind, or an old project that could be updated with transistors or integrated circuits, let me know about it; I'll pass the idea along to one of our authors.

We are continually trying to improve **ham radio,** but we can only do that successfully if we hear from you, the reader, so keep those cards and letters coming in.

Jim Fisk, W1DTY
editor



solid-state receiver for 160 meters

Although this receiver was designed for the 1.8-MHz band, it contains many interesting circuits that could be applied to higher frequency units

John K. Gotwals, W3TNO, 419 West Stadium Avenue, West Lafayette, Indiana, 47906

Recent changes in the FCC rules governing amateur activity on 160 meters have prompted renewed interest in 160-meter operation. Since many ham-band-only receivers do not cover 160, I felt there should be a lot of interest in a high quality, solid-state receiver capable of tuning 160 meters. The receiver presented here uses many of the latest semiconductor devices, and many of the design ideas and circuits would be equally applicable to higher-frequency receivers.

A block diagram of the 160-meter receiver is shown in **fig. 1.** Note that there is no rf stage ahead of the mixer. This omission is not an attempt to simplify the receiver but to improve its over-all performance. Several years ago recent receiver developments were reviewed,[1] and the authors pointed out that to minimize spurious responses, no rf stage is needed, or desired, at frequencies below 15 MHz.

In the receiver shown in **fig. 1** a 455-kHz i-f was chosen so I could try the Murata CF-455 ceramic filters.[2] These filters are compact, exhibit good shape factors and are easy to use with simple resistance-coupled transistor circuitry. Collins mechanical filters could be used as well.

variable attenuator

A schematic of the variable attenuator and high-pass filter is shown in **fig. 2.** The attenuator can be varied from zero to 40 dB in 10-dB steps[3] and is used if you are listen-

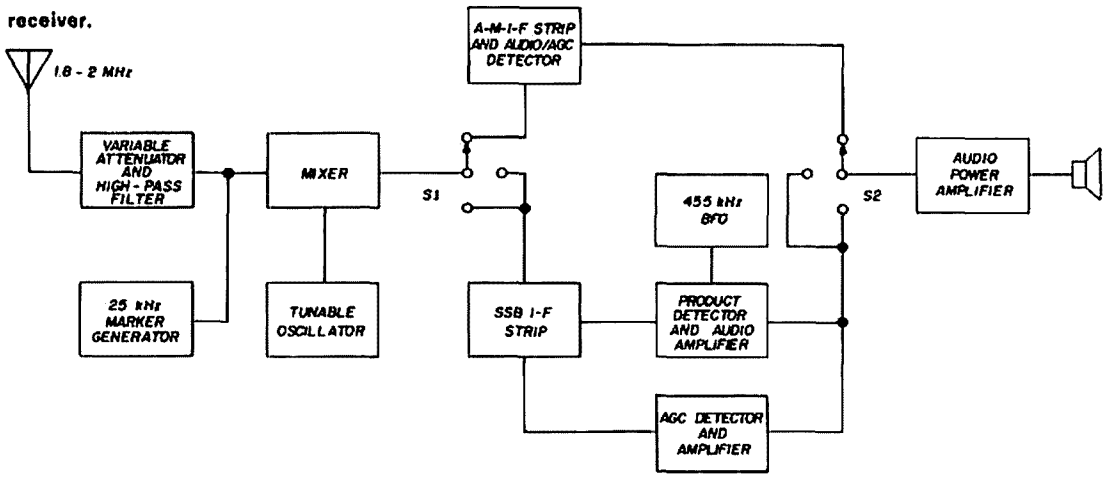
ing to local stations or if you want to drop out weak signals and hear only the strong ones.

The high-pass filter is used to eliminate cross modulation and intermodulation from local broadcast stations. Although the mosfet mixer has excellent square-law characteristics, you still need some front-end rejection of the extremely strong broadcast signals present on a long 160-meter long-wire antenna.

down corner frequency is 1800 kHz. At 1604 kHz, response is down 45 dB. An excellent discussion and application of modern network synthesis to the design of low- and high-pass filters is given by Wetherhold.[4] The serious filter designer will also take a look at the work by Geffe.[5]

I used 5 percent mica capacitors in my filter, but performance closer to the theoretical would be obtained if more expensive 2 percent mica capacitors were used. Both

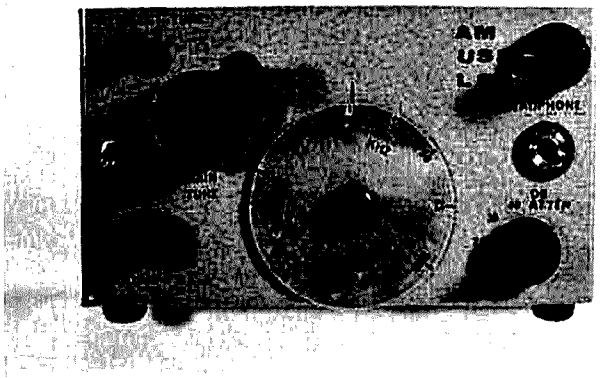
fig. 1. Block diagram of the solid-state 160-meter receiver.



The high-pass filter is an elliptic function type with response zeros at 918 kHz, 1586 kHz and 1415 kHz. The zeros are the frequencies where the theoretical attenuation becomes infinite. The theoretical passband ripple is less than 0.2 dB, and the 0.2 dB

types are available from Allied Electronics. The Q of the inductors should be at least 100; I used Ferroxcube 4C4 pot cores. They have the advantage of being adjusted approximately ±6 percent. In addition, they have a negative linear controlled temperature coefficient that compensates silver-mica capacitors. They cost about $2.50 apiece in the 14 mm size. Write to Ferroxcube[6] for more information.

Receiver front panel.



dual-gate mosfet mixer

A schematic of the mixer front end is shown in fig. 3. The data sheet for the 3N141 shows a mixer biased to 3 mA with both gates at about the same potential. In a companion application note,[8] H. M. Kleinman mentions two requirements for spurious response elimination: clean oscillator waveform and properly adjusted operating conditions and signal levels. This means that biasing voltages are important. Also, local-oscillator injection should not be too high.

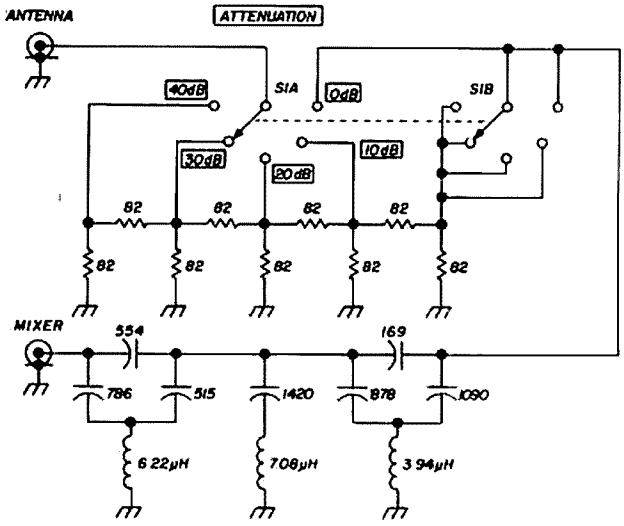


fig. 2. Variable attenuator and high-pass filter for the solid-state receiver. Input and output impedance of the filter is 52 ohms. With Ferroxcube 1408C-A25-4C4 pot-core assembly, 1408HD hardware and 1408F1D bobbin, 13 turns no. 24 enameled was usod for the 3.94 μH Inductor, 16½ turns for the 6.22 μH inductor, and 17 turns for the 7.08 μH inductor. Higher Q could be obtained by using 19/46 Litz wire.

As a compromise between conversion gain and spurious-response generation, I used a 2.5 volt p-p signal to the mixer.

I used electronic tuning for the input circuit so I would have complete freedom in the placement of the mixer module. If you're interested in using tuning diodes in the place of mechanical tuning, Motorola has two helpful application notes.[9,10]

local oscillator

The stability of the receiver is determined by the local oscillator, so circuit design is very important. As noted above, the stage should operate in class A for minimum spurious response. Fortunately, class A operation is also important to frequency stability. In simple oscillators an important source of frequency instability is associated with the intermodulation of harmonic frequencies which are produced by the limiting that takes place if agc is not used.[11] The Clapp oscillator in this design has been linearized by adding an agc circuit; frequency stability, with variable supply voltage, is improved by about 15 times.

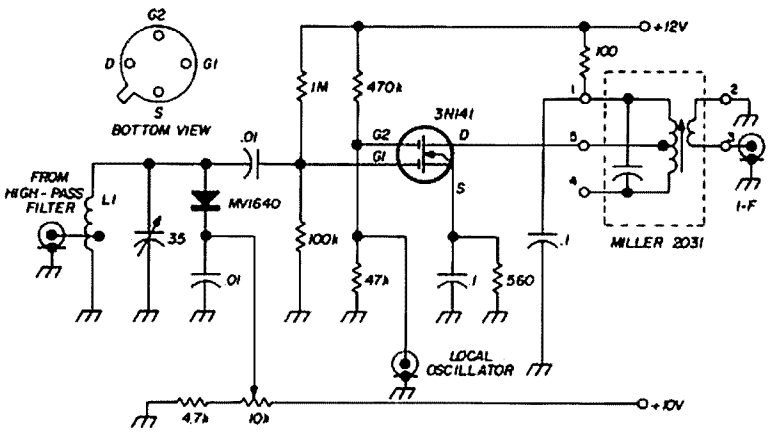
The oscillator circuit is shown in **fig. 4.** Diode D1 linearizes the oscillator as follows:

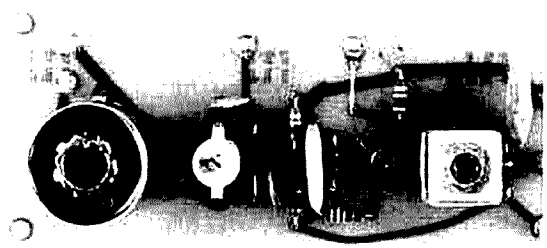


High-pass filter uses high-Q pot-core coil forms.

when power is applied, the gate is at the same potential as the source so forward transconductance is maximum. As oscillations build up, D1 clamps the gate below ground; this charges up capacitors C1 and C2 and results in a negative dc voltage at the gate, and the forward transconductance of the device is lowered until equilibrium is reached.

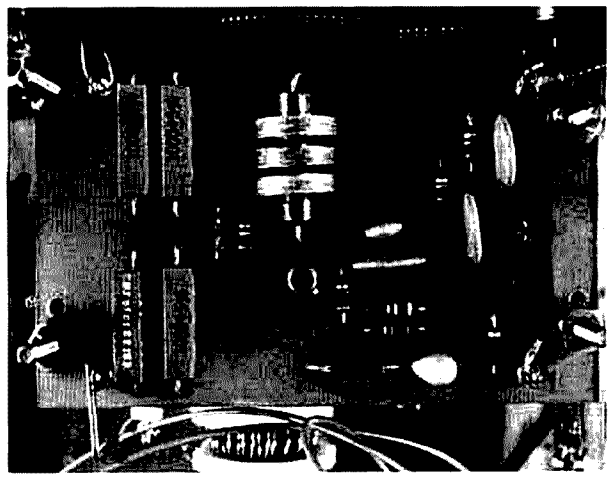
fig. 3. Dual-gate mosfet mixer. L1 is 115 μH and consists of 41½ turns no. 24 on Ferroxcube 2213C-A60-4C4 pot core, tapped 1 turn from cold end.





Without D1 in the circuit, the 3N152 would have to provide its own limiting, and this could only be done with the mosfet operating in a nonlinear mode. W2YM uses a diode in a similar circuit and reports that it stabilizes the oscillator.[12] Capacitor C3 is selected to provide the desired injection level at the mixer.

For best operation, the Q of L1 should be high. I used two Micrometals T94-2 toroid cores in series.* The tuning capacitor is from a surplus ARC-5 transmitter. This capacitor is an excellent choice for any vfo. Since the capacitor is quite rigid, I mounted all the vfo components on the capacitor frame. The gears that are an integral part of the ARC-5 capacitor are used for the dial.



Vfo board is mounted on top of the ARC-5 tuning capacitor. The series toroidal coils ore mounted on a ¼" brass bolt which is soldered to the capacitor frame. Teflon pads are used to clamp the coils.

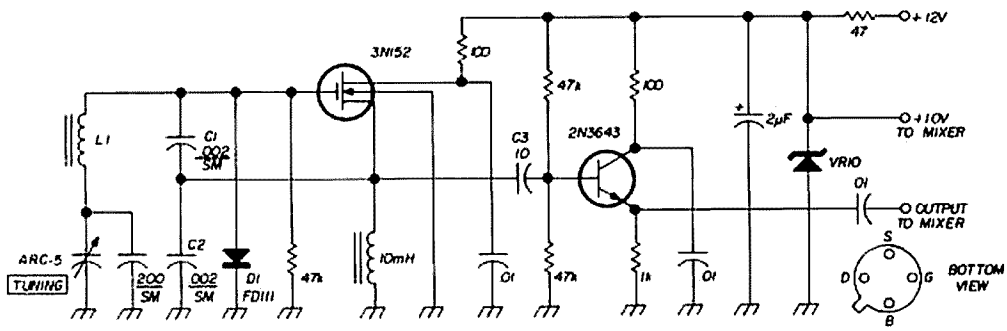


fig. 4. High-frequency oscillator. L1 is two Micrometals T94-2 toroids in series, 41 turns no. 22 on each.

a-m i-f strip and detector

The a-m i-f strip shown in fig. 5 uses a unique integrated circuit, the LM372 manufactured by National Semiconductor.[13] This IC is an i-f amplifier, a-m detector and an audio amplifier with self-contained agc. This gem will take an amplitude-modulated rf signal at frequencies up to 2 MHz and give you 0.8 V p-p audio output. The agc provides an essentially constant output as the rf input varies over the range from 50 microvolts to 50 millivolts.

In the photo of the a-m i-f strip the 3- and 6-kHz Murata ceramic filters are on the left. The 6-kHz filter is used with the a-m i-f system, and the 3-kHz filter is used for ssb. Resistor R1 is adjusted so that maximum

*Micrometals toroid cores are available from Amidon Associates, 12033 Otsego Street, North Hollywood, California 91607.

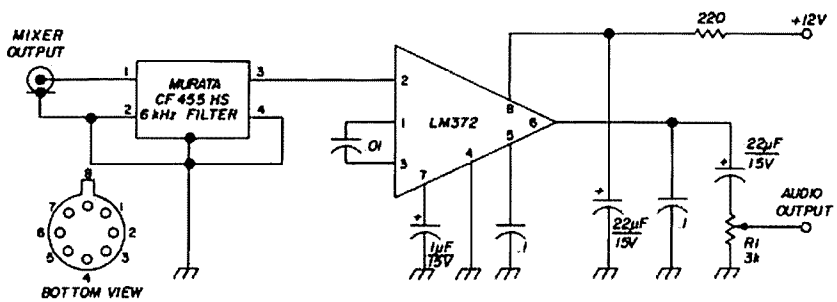
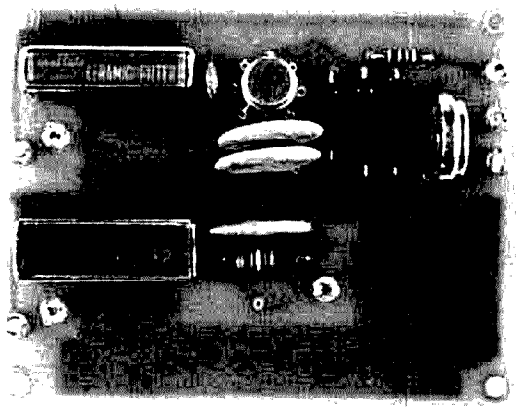


fig. 5. A-m i-f strip uses a communications-type integrated circuit, the National Semiconductor LM372. R1 is adjusted for 200 mV output.



A-m i-f board.

MHz. However, since this is a "first-generation" IC, agc detection and a ssb product detector must be furnished externally.* One of the big advantages of using the LM371 is that you don't have to worry about internal feedback. No neutralization is needed, even at 100 MHz. However, proper shielding must be used between input and output.

The LM371 data sheet shows the input and output impedances of the device over its complete operating range. However, for my application on the relatively low frequency of 455 kHz, I didn't worry about a

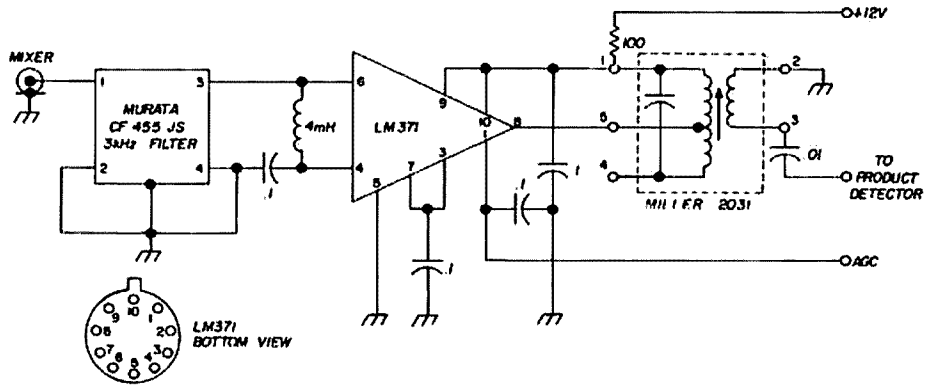


fig. 6. Single-sideband i-f system requires an external agc input from the agc circuit shown in fig. 9.

audio output is 200 mV (200 mV drives the audio-output stage to full output).

ssb i-f strip

The ssb i-f strip shown in **fig. 6** is based on the use of an LM371 integrated circuit. The LM371 is quite versatile and may be used as an amplifier from dc to over 100 perfect match, and I used readily available i-f transformers.

*A newer device, the National Semiconductor LM373 i-f strip, is a broadband communications IC capable of performing the diverse functions required in a-m, fm and ssb receivers and transmitters. Simple external connections convert the i-f strip from one mode to another. The LM373 has a self-contained audio agc system with fast attack and slow release.

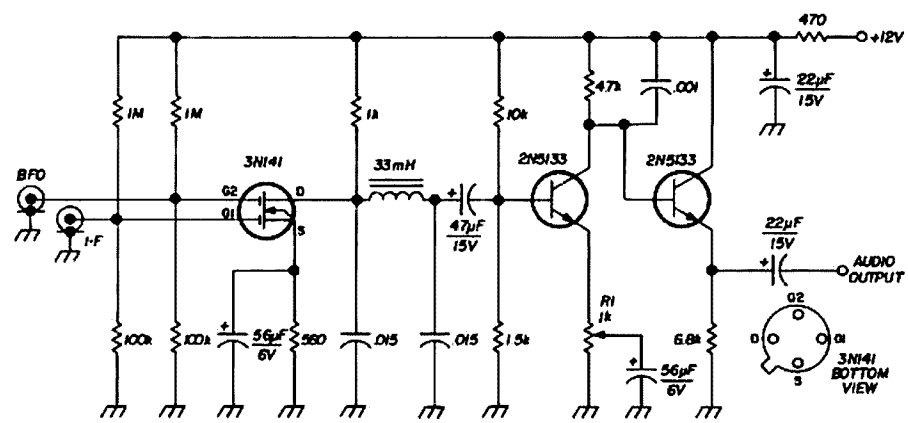


fig. 7. Product detector and audio amplifier.

product detector

A schematic diagram of the dual-gate mosfet product detector is shown in fig. 7. An adjustable-gain audio amplifier is coupled to the output of the product detector and is used to provide the same audio output as the a-m i-f system. Resistor R1 is adjusted so that the gain of the ssb i-f strip (with no agc) is the same as that of the a-m i-f strip. The product detector is biased the same as the high-frequency mixer —at 3 mA drain current. This product detector has very low intermodulation distortion and actually performs much better than the diode type in my commercial receiver. Let me hasten to add, though, that diode product detectors are capable of excellent performance if they are properly designed.[14]

The correct crystal frequencies are determined by the *i-f* filter. Although most single-sideband operation on 160 meters is confined to lower sideband, now and then you hear a station on upper sideband, so it pays to be able to receive either.

The diode in the gate circuit of the 2N4360 oscillator linearizes the oscillator and results in an extremely clean waveform. The 2N5163 source follower isolates the oscillator stage from any loading by the product detector. Resistor R1 is selected for 2 V p-p at the gate of the product detector.

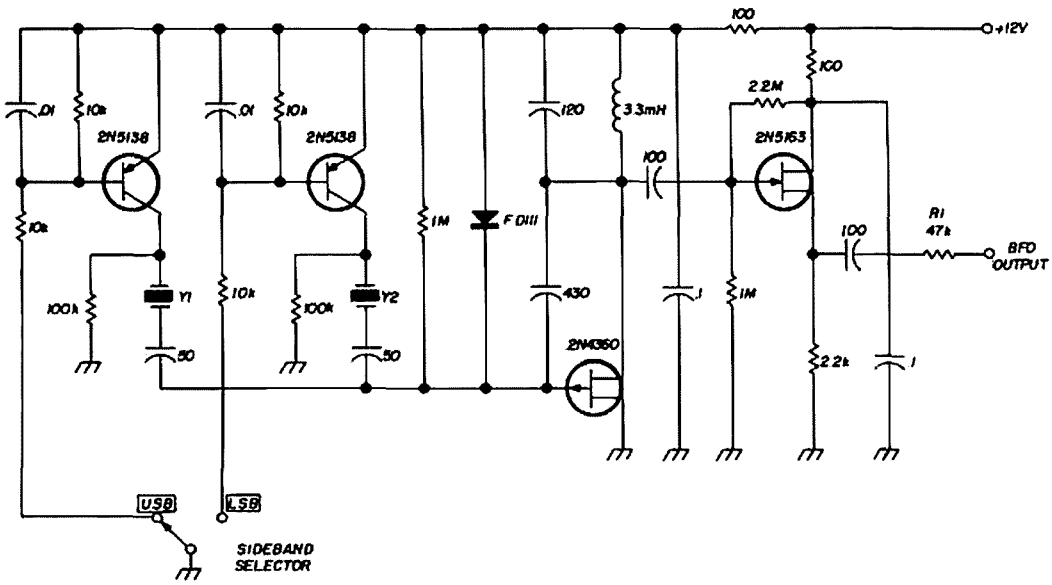
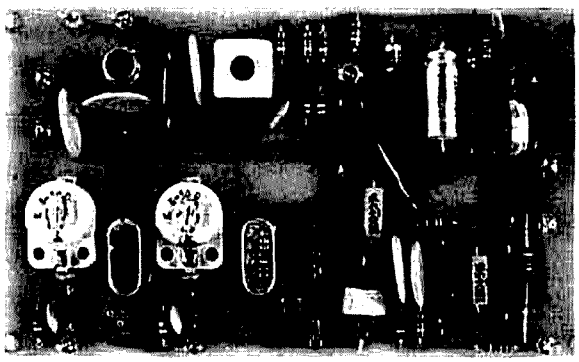


fig. 8. Beat-frequency oscillator uses electronic crystal switching. R1 is selected for 2 V p-p at the gate of the product detector.

beat-frequency oscillator

The crystal controlled bfo shown in fig. 8 uses electronic crystal switching. The circuit could be simplified if a mechanical switch was used. However, it seemed to me that the flexibility of chassis layout outweighed the disadvantage of two extra 2N5138 transistors and a few extra resistors and capacitors.

Bfo, product detector and audio amplifier board.



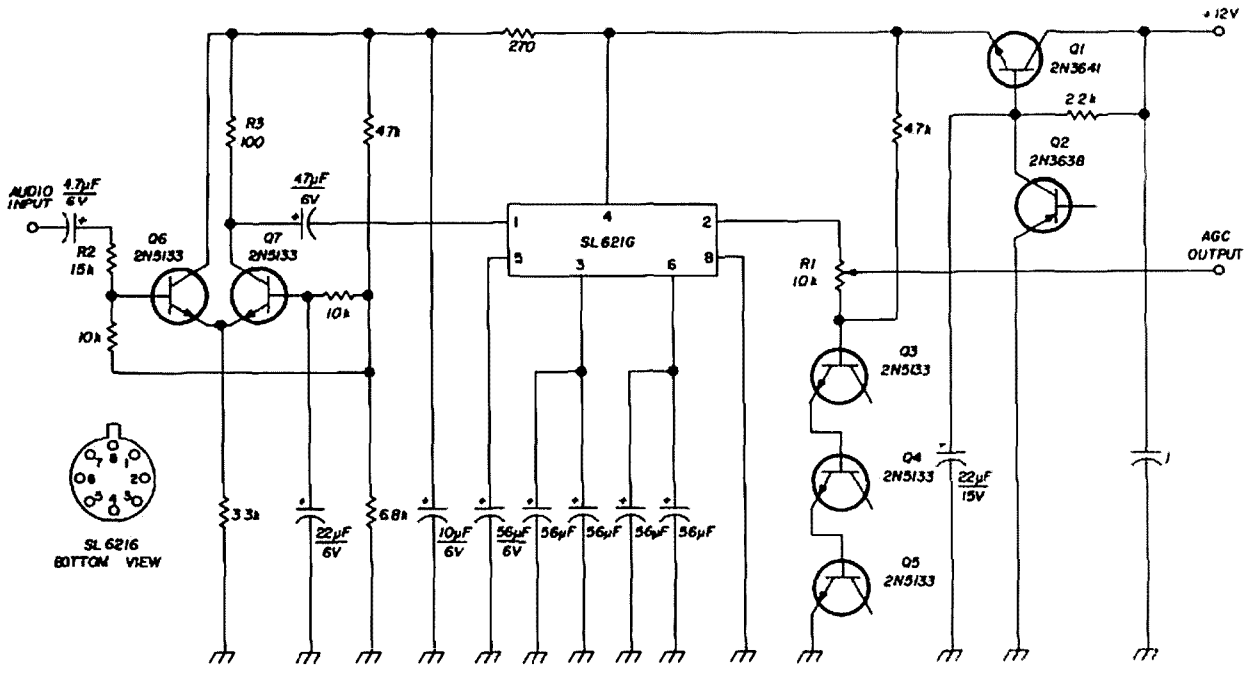


fig. 9. This versatile ssb agc circuit is based upon a communications integrated circuit, the Plessey SL621G.

The i-f amplifier, product detector, low-level audio and bfo are all mounted on the same board as shown in the photograph. However, in retrospect I would recommend that the i-f amplifier and its associated components be mounted on the same board as the i-f ceramic filter. In my layout the bfo radiates into the input circuit of the i-f amplifier. Although this doesn't appear to cause any trouble, moving the i-f amplifier to another board would prevent it from happening.

agc detector and amplifier

Good ssb and cw agc circuits are hard to find, although it's easy enough to list the kind of performance we're looking for:

1. Fast attack, so the first word spoken doesn't blast out of the speaker.

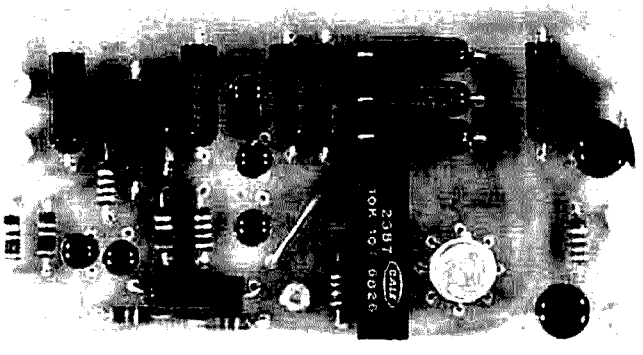
2. Slow release—you don't want to hear the background noise between speech pauses—but not so slowly that the agc won't follow normal signal fading.

3. Immunity from noise interference.

The problem in designing a good agc system arises from the conflicts between the three basic performance requirements. The literature has a number of "hang" agc circuits—and they sure hang up on noise! Circuits can be designed that will pretty well meet the three basic requirements, but if they're built with discrete components, they are quite complex.

The Plessey SL621G[15] is an agc generator which comes pretty close to meeting the standards. It has fast attack and slow release for speech. By using two detectors, one with short rise and fall times, and one with long rise and fall times, the SL621G

Ssb agc board.



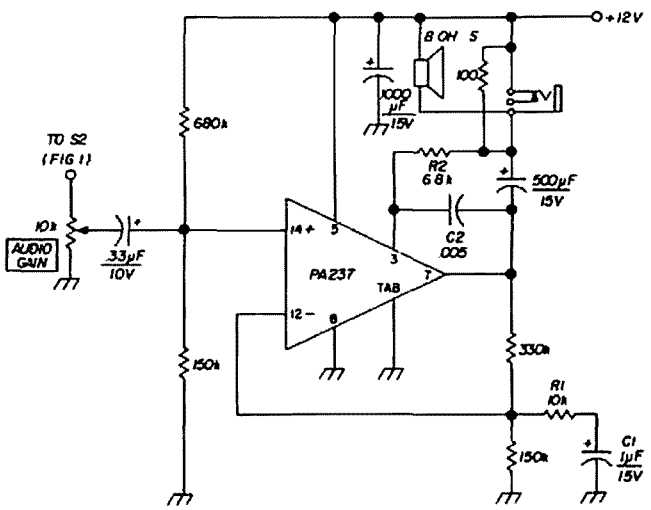
is able to prevent short noise pulses from "hanging up" the agc. If you are listening to a clear channel, and a noise spike comes along, the agc is activated only for the duration of the noise burst by virtue of the short-time-constant detector; the long-time-constant detector "remembers" the pre-noise agc level.

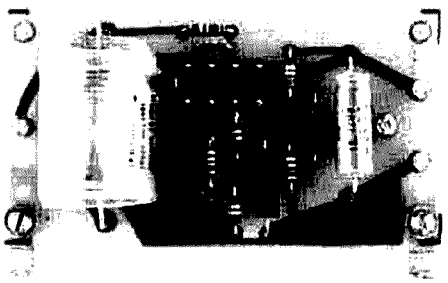
The same thing happens if a noise burst is superimposed on a speech signal. The short-time-constant detector initiates rapid agc action; when the noise pulse is over, the agc just as rapidly returns to the pre-noise level set by the long-time-constant detector.

A schematic of the complete agc module is shown in **fig. 9**. Don't be scared by its apparent complexity—2N5133 transistors are only 17¢ apiece. The circuit would be a lot more simple if I had used the other members of Plessey's SL600 series integrated circuits, but unfortunately industrial rated units (G suffix) have only recently become available.

Transistors Q1 and Q2 are used to reduce the 12-volt supply to the 6-volt level required by the SL621G. Q3, Q4, Q5 and resistor R1 make the agc output voltage compatible with the LM371 i-f amplifier. These components ensure that the agc output tracks the internal LM371 agc reference voltage as the temperature changes. Transistors Q6 and Q7 are connected as an emitter-coupled clipper and prevent the input coupling capacitor to the SL621G from charging up on strong audio peaks. Resistors R2 and R3 reduce the 200 mV input audio signal to the 10 mV level needed by the IC.

fig. 10. Audio output stage provides up to 1 watt into an 8-ohm speaker. Note that phone-jack frame is isolated from the chassis.





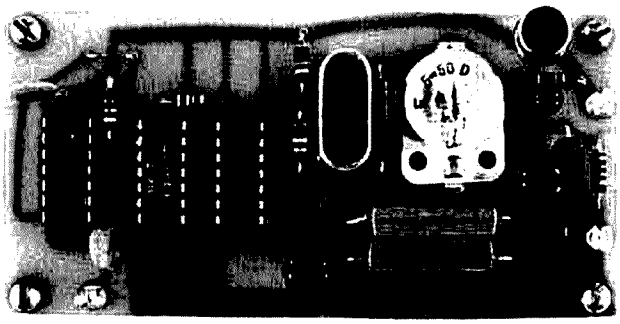
Audio power stage uses PA237 integrated circuit.

audio output stage

The audio output stage shown in **fig. 10** is driven to full output with a 200 mV input signal (full output is about 1 watt into an 8-ohm speaker). The output audio has very low distortion because of the generous amount of negative feedback that is used in the circuit. Resistor R1 sets the amount of feedback, the R1-C1 series circuit determines low-frequency cutoff. High-frequency cutoff is determined by R2-C2.

Resistor R2 also sets the quiescent current drawn by the quasi-complimentary output of the PA237 IC. Since R2 is connected to the positive 12-volt supply (through the speaker) it protects the integrated circuit in case the speaker is not connected. Because of the positive feedback (load bootstrapping) that results when R2 is connected this way, the output has a larger voltage swing.

The complete audio output circuit board is shown in the photograph. Note that the 1000 µF, 15 V capacitor shown in **fig. 10** is not on the board. I mounted it adjacent to the audio output module. This capacitor is needed for the current demands of the PA237 on audio peaks.



Crystal calibrator board.

marker generator

Since the 160-meter band is divided into 25-kHz segments, band markers every 25 kHz are a definite operating aid. I decided to use a 1 MHz crystal because it is more stable, smaller and less expensive than a 100 kHz unit. However, the cost of an extra divide-by-10 counter partly cancels some of the advantages. I opted for stability and used the 1 MHz approach.

The complete calibrator is shown in **fig. 11.** Transistors Q1, Q2 and Q3 reduce the 12-volt supply to 5 volts; Q4 is the 1 MHz oscillator. Since 25 kHz markers are desired, the crystal frequency must be divided by 40. This is done in two steps: the MC853P dual flip-flop divides by 4; then a Fairchild CμL9958 divides by 10. The square-wave output of the 9958 drives a TTμL9601 one shot—this provides a 50 nanosecond pulse each time it is triggered.

The 50 ns pulse has rise and fall times of better than 10 ns, so it is rich in harmonics. Unlike calibrators with square-wave outputs, adjacent harmonics from a pulse generator have equal signal strength.[16] Capacitor C2 is chosen to provide markers of the desired signal strength at the input to the receiver.

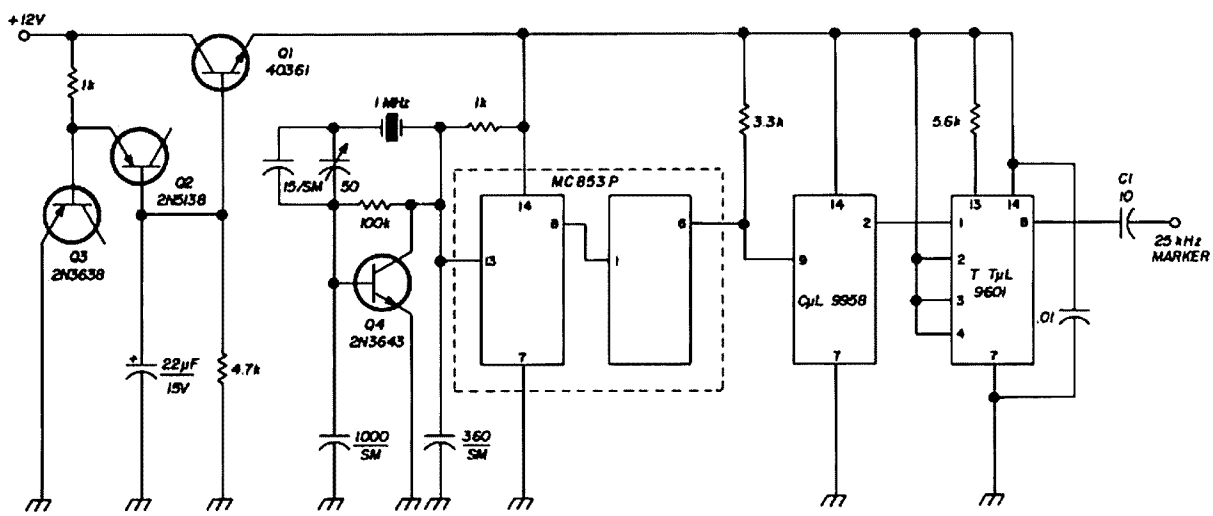
power supply

The 12-volt power supply for the receiver shown in **fig. 12** furnishes up to 400 mA at 12 volts. Since the audio output stage draws quite a bit of current during audio peaks, a stiff power supply is needed to prevent feedback between the audio-output stage and the other circuits through the power supply. The heart of this power supply is the Fairchild μA723C voltage regulator IC. This IC has a built-in voltage reference of about 7 volts that is extremely stable with changes in temperature. The remainder of the IC functions as an operational amplifier.

As shown in **fig. 12,** the voltage reference, pin 4, is connected to the non-inverting input (pin 3). The inverting input (pin 2) is furnished with a voltage by R1 that is proportional to the output voltage. The operation of a power supply of this type has been discussed in **QST.**[17]

One feature of this power supply is the

fig. 11. Marker generator provides accurate 25-kHz markers from a 1 MHz crystal-controlled oscillator.



use of foldback current limiting. With most short-circuit-proof power supplies, the dissipation of the series-regulating transistor Q1 is quite high when the output is shorted to ground. This is because the voltage across the transistor is equal to the supply voltage, and the current supplied by Q1 is the full-load value. Thus, the power dissipated by Q1 during an output short circuit is three to four times higher than the power

R2 and R3 should be wirewound types. The complete power supply is shown in the photo.

summary

I hope this article will encourage more hams to build, and not buy, their next piece of equipment. Once you take the initial step, designing and building ham gear can be fun. With transistors and multi-function

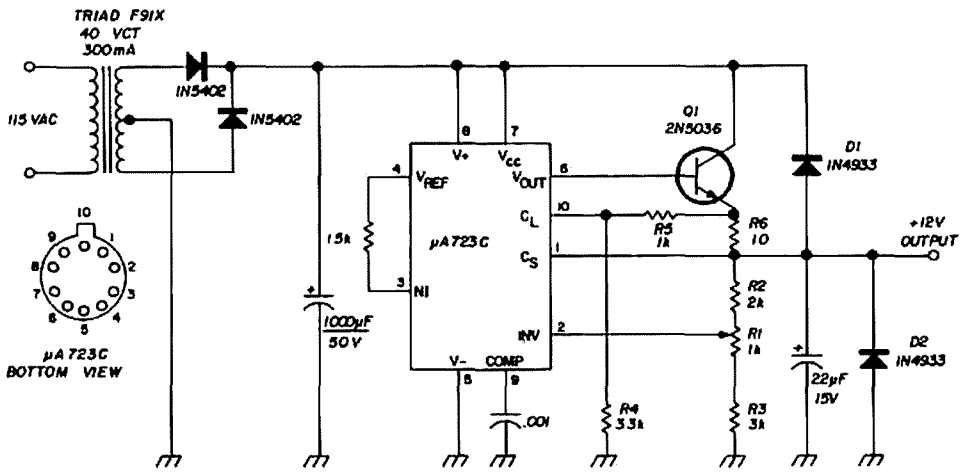


fig. 12. Regulated power supply uses a voltage-regulator IC in a foldback-current-limiting circuit. Output current is limited to 400 mA.

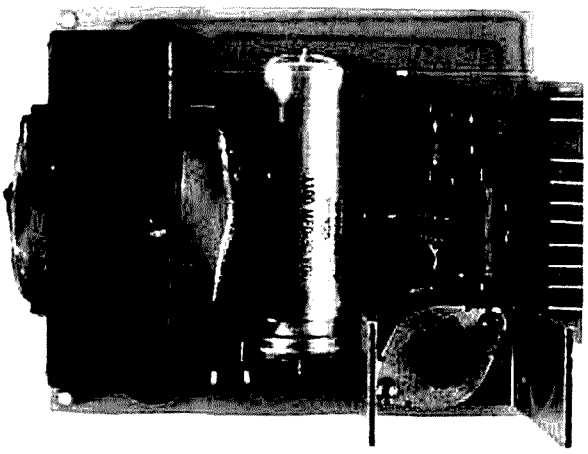
dissipated under normal full-load operation. This means that you have to use a larger heat sink to protect the transistor during long-duration short circuits.

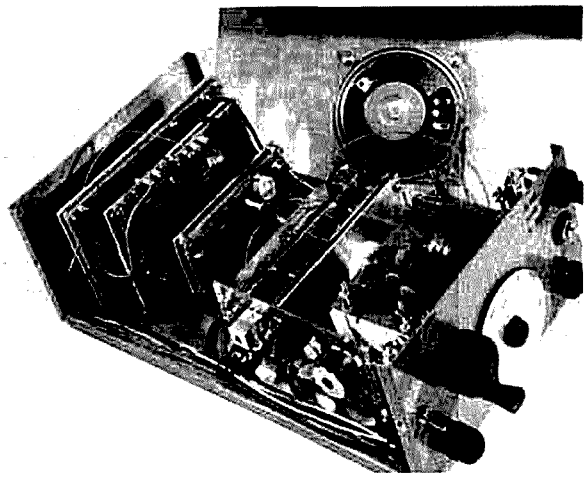
However, in the power supply in **fig. 12** the current through Q1, when the output is short circuited, is only 85 mA. In normal operation the output voltages remains at 12 volts until the output current reaches 400 mA. If the load resistance decreases further (load current increases) the voltage and current decrease to their short-circuit values of 0 volts and 85 mA respectively; this is called foldback current limiting. Resistors R4, R5 and R6 set the current knee and short-circuit current. The μA723 data sheet provides a formula for selecting resistor values for different applications.

Diodes D1 and D2 are protective diodes that are useful in any power supply[18]. For stable power supply operation, resistors R1,

integrated circuits, it is possible to come up with circuits that would have been impractical a few years ago. And in the future we can look forward to even more exciting new devices.

Power supply uses feedback current limiting.

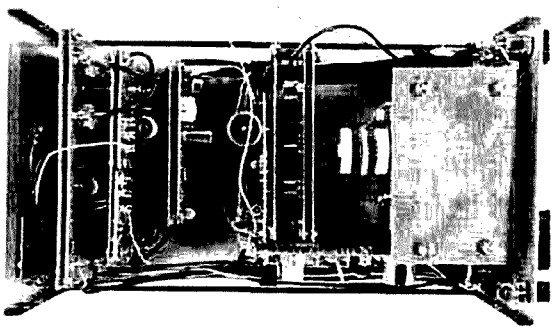




Internal construction of the receiver. Board positions from the rear as follows: bfo, product detector and audio stage; a-m i-f strip and filters, and ssb agc; 25-kHz calibrator; audio output; high-pass filter; and vfo. The mixer board is mounted on the side of the tuning capacitor.

Bob Hirschfeld, W6DNS, who designed the LM372, has stated that, "future developments in the communications IC area are going to raise a few more eyebrows."[19] Engineers at Fairchild Semiconductor make similar statements: "Second generation linear circuits will make their biggest impact in the field of communications . . . An entire family of circuits is under development for manpack mobile, and ground-based communication system."[20]

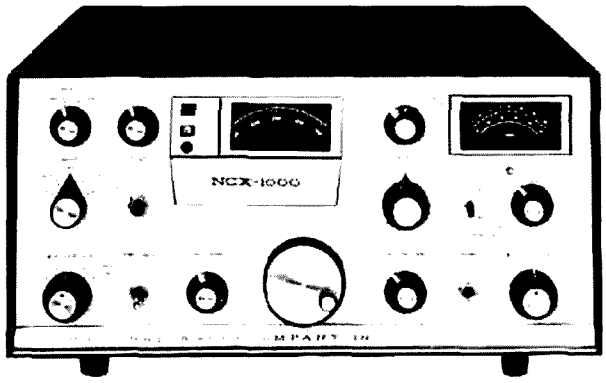
I wish to thank Don Coppage for taking the photographs of the complete receiver.



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ham radio



NCX-1000
1 Kw Solid-State Transceiver
(80 thru 10 Meters)

Here's a transceiver designed for the amateur who would rather spend his hard-earned radio dollar on performance than frills. The NCX-1000 is built to meet the demands of the operator who needs and desires a high performance SSB-AM-CW-FSK rig with solid-state dependability and plenty of power. Add to this the convenience of having your transmitter (including linear amplifier), receiver, power supply, and monitor speaker in a single, compact, smartly styled 59 pound package.

So let's look at the NCX-1000, starting with the double-conversion, solid state receiver. After the received signal is processed by a double-tuned preselector, a stage of RF amplification, and another preselector, it is applied to the first mixer for conversion to the first IF frequency. The first IF contains passband filters and a stage of amplification. A second mixer then converts the signal to the second IF frequency for additional processing by a 6-pole crystal-lattice filter and four IF stages. Finally, the signal is detected and amplified by four audio stages. The unparalleled high dynamic range lets you tune in weak stations surrounded by strong interfering signals. The result? High performance for SSB, AM, CW, and FSK. Sensitivity of 0.5 EMF microvolt (for a 10 db S+N/N ratio).

In the transmitter you'll find three stages of speech amplification followed by a balanced modulator, a crystal-lattice filter, a filter amplifier, and an IF speech processor (clipper). A mixer converts the signal to a first IF frequency for processing by two crystal passband filters, and two IF amplifiers. A second mixer converts the signal to the transmitting frequency where it is amplified in five RF stages before it gets to the grid of the 6BM6 driver. Final power amplification takes place in a forced-air-cooled 8122 ceramic tetrode which feeds the antenna through a pi network. Other features? You bet! Grid block keying for CW. Complete metering. Amplified automatic level control (AALC).

So here's a package that can give you 1000 watts PEP input on 80 through 10 meters, 1000 watts on CW, and 500 watts for AM and FSK. The speech processor lets you double your SSB average power output with minimum distortion. No frills with the NCX-1000. Just top performance.

Write for complete details.

NATIONAL RADIO COMPANY, INC.
NRCI

111 Washington Street, Melrose, Mass. 02176
617-662-7700

low-band converted-vee antenna

This trapped antenna provides low swr operation on 40, 75 and 80 meters—great for 5BDXCC

Bob Polansky, W6JKR, 1024 Green Lane, La Canada, California 91011

With the new 5-band DXCC award, interest in 40 and 80 meter operation has vastly increased. A number of W/K stations have completed, and many others are well on their way, to making the required contacts with one hundred countries on each of these bands. One could hardly avoid becoming intrigued with this challenge. Since my receiver and transmitter were already capable of covering these bands, my efforts turned toward a suitable antenna system.

dual-band inverted vee

With a central mast on my property and little chance of erecting other high poles the best compromise antenna appeared to be a combination 80- and 40-meter inverted vee patterned after the dipole described by Neil Handel, W1IR.[1] The 80 and 40 meter elements were separated with 1/4 x 6 inch wooden-dowel spacers simmered in a high-melting-point wax, as W1IR had done, for weather protection.

Since I operated for long periods of time on either cw or ssb I tuned the long legs of the inverted vee to 3.8 MHz for ssb operation and made provisions for connecting across egg insulators with shorting devices to lengthen each leg for 3.5 MHz cw operation. That way a relay and its associated wiring was avoided.

Clip wires could be used for shorting devices; however, they suffer several disadvantages. First, they can be inadvertently pulled off by branches or other physical obstructions. Second, it is possible that the clips may make poor electrical contact because they're connected onto a corroded spot on one of the wires. My solution was to use a shorted 83-1SP plug inserted into a receptacle soldered across an egg insulator.

The first combination antenna I used

is shown in **fig. 1**. The apex of this inverted vee was 55 feet above ground and the included angle between the antenna legs was approximately 120°. Length l_1 was chosen for the best compromise swr over the 7 MHz band, favoring the cw end; l_2 was chosen for the best compromise swr in the 3.8 MHz area, also favoring the lower ssb frequency with the shorting plugs removed; l_3 was chosen for the best compromise swr in the 3.5 MHz area again favoring the lower cw frequency with the shorting plugs in place. Note that these lengths will vary somewhat as a function of antenna apex heights, end heights and included angle.

MHz. The frequencies between 3.6 and 3.8 MHz were of no concern for transmitting purposes. With this in mind, my thoughts immediately turned toward a parallel LC trapped antenna scheme. Would it be possible to develop, with reasonably sized components, a trap with high enough Q to be effective at 3.8 MHz and not at 3.5 MHz? If it were, the wire inside the trap (toward the center of the antenna) would load well in the 3.8 MHz region, while the inductive reactance added by the trap itself plus a short length of wire outside the trap might load in the 3.5 MHz region. The idea sounded crazy enough to work.

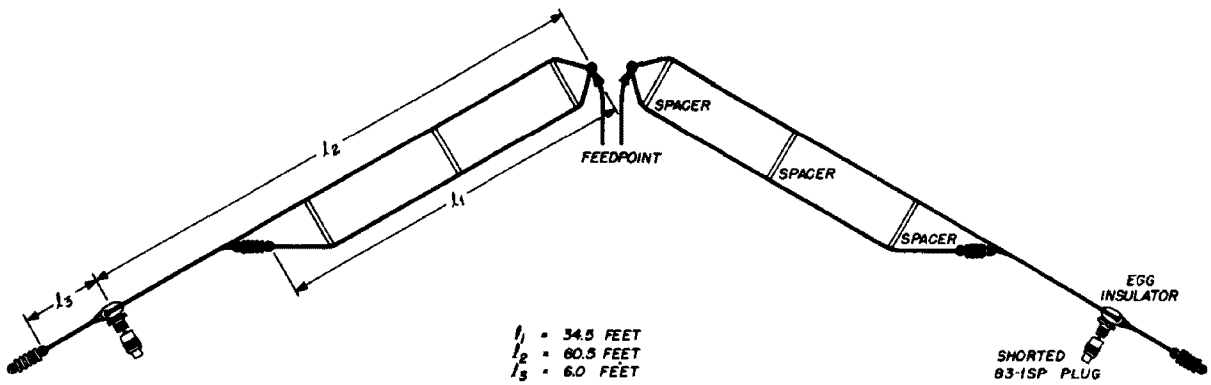


fig. 1. Initial 80-, 75- and 40-meter antenna system.

No interaction was noted between the 40- and 80-meter elements of this antenna system while experimenting with the lengths of each. The indicated swr for 40-meter operation was always less than 2:1 while on 80 meters it was possible to cover 300 kHz with an swr less than 2:1.

I soon found myself making numerous trips to the roof to remove and install the shorting plugs as exotic DX stations kept appearing in the 80 meter sub-band for which I was **not** set up. Certainly, I thought to myself, another solution, one not needing relays, clip wires, or other physical changes to the antenna system was possible.

converted vee

For planning purposes, operation was desired at an swr of less than 2:1 between 3.5 and 3.6 MHz, and between 3.8 and 4.0

trap construction

Recent issues of the ARRL Radio Amateur's Handbook discuss the construction of such traps. Basically, it involves the following:

1. Two loops of the antenna wire you intend to use are passed through an egg insulator in the normal manner. Both ends are then wrapped, cut off close to wrap, and soldered, leaving the insulator with two loops of wire.

2. The tuning capacitor is soldered across the two independent loops.

3. A larger than necessary inductor is soldered across the insulator-capacitor combination.

4. The entire assembly is grid dipped, and the inductance trimmed until the dip occurs at the proper frequency.

5. The LC circuit is then disassembled, the actual antenna wire wrapped around the egg insulator as before, and the trap rebuilt and soldered.

The traps I built used 50 pF Centralab ceramic transmitting capacitors, part number 850S-50Z,* which are rated at 7500 Vdc. A greater safety margin could be obtained by using 15,000 Vdc capacitors, but the cost would have increased by a factor of more than three.

For the inductor, 2½-inch OD lucite tubing 5 inches long was used as a coil form. The form was slotted on a lathe at 10 turns per inch. To resonate the traps at 3.805 MHz, the design frequency, the required inductance was 33.7 microhenries. This was fabricated by winding 33 turns of number 14 wire on the preslotted lucite form. The LC combination was then fine-tuned to the proper frequency by trimming the inductance and observing the results on a well calibrated grid dip meter. The final coil ended up being 30 turns.

Another reasonable approach is to buy a 10 inch Air Dux coil, which is available from WRL. Either of two coils would suffice: the Air Dux 2010T (2½-inch diameter, 10 turns per inch number 16) or the Air Dux 2410T (3-inch diameter, 10 turns per inch number 14). These coils, when split in half, should provide more than enough inductance to build two traps.

A note of caution is in order here. When trimming the coils, monitor the grid-dip oscillator output on a well calibrated receiver—don't trust the oscillator's calibration. The grid-dip oscillator dial I used was more than 100 kHz out of calibration.

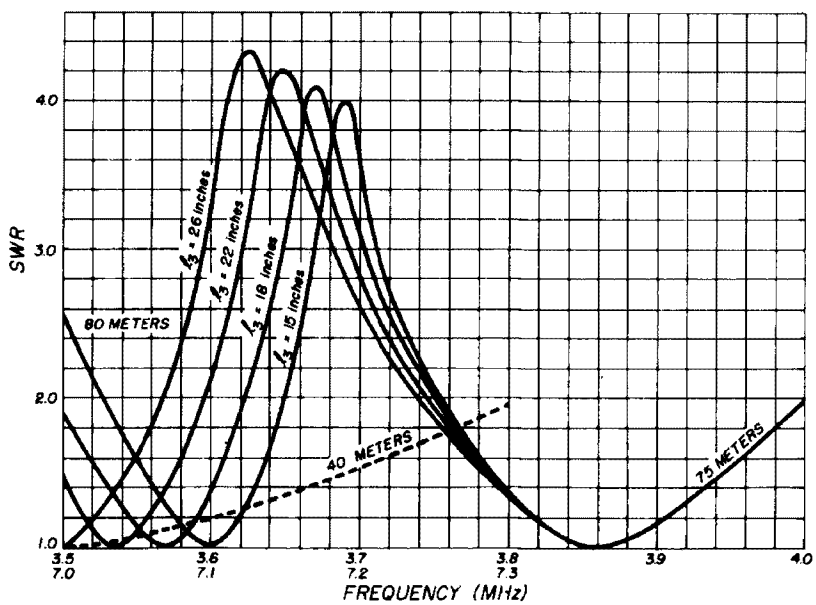


fig. 2. Standing-wave ratio performance of the converted-vee antenna.

weatherproofing

Since hollow lucite forms were used for my traps, end caps were fashioned on a lathe and cemented into the hollow ends of the forms to keep excessive moisture out of the soldered components of the traps. Drain holes were drilled in the low end of the traps to let trapped moisture drain out.

The traps have not yet been subjected to the rainy season, so I don't know how the wire wrapped around the outside of the forms will hold up. If Air Dux coils are used, a soft, flexible plastic container of some type should be used around the traps.

* The 50-pF Centralab 850S-50Z ceramic transmitting capacitors are available from Allied Electronics, 100 N. Western Avenue, Chicago, Illinois 60680. Order part number 43D1101, $2.70 each plus shipping.

performance

With the traps installed in both legs of the antenna in place of the shorting plugs various lengths of l_3 were tried. Success! Two nulls in swr were noted; one in the 3.8 MHz region and one around 3.5 MHz. Changing l_3 had little effect on the 3.8 MHz swr characteristics; however, it did move the swr null across the 3.5 to 3.6 MHz region, as shown in **fig. 2**. An l_3 of 18 inches produced an swr of less than 2:1 for a range of frequencies from 3.5 to 3.64 MHz.

The experimenter may wish to modify the characteristics of either the 7, 3.8, or 3.5 MHz portions of the antenna system. Since no interaction between the 80 and 40 meter elements was observed, each can be trimmed independently. Shortening l_1 to 33 feet moved the swr null to 7.3 MHz. Intermediate lengths will produce swr nulls at other frequencies in the 7 MHz band.

For the chosen 3.8 MHz parameters, the 3.5 to 3.7 MHz characteristics can be modified by selecting different lengths of l_3, as indicated in **fig. 2.** If a change in the 3.8 MHz characteristic is desired, both l_2 and the trap frequency will have to be changed simultaneously. These changes will undoubtedly modify the length l_3 required to achieve the same 3.5 MHz characteristic as before. No actual data were obtained on this.

One interesting observation should be noted here. The traps in my antenna were tuned on the bench to 3.805 MHz. When checked, installed in the antenna, their resonant frequency had lowered to 3.760 MHz. This was expected, due to the added inductance of the antenna; however, the swr null at 3.860 MHz came as a pleasant, but unexpected surprise.

The finished converted-vee antenna is shown in **fig. 3.** Note that the apparent capacitor value in each trap is 52 pF (50 pF from the capacitor itself and 2 pF from the antenna wire wrapped around the egg insulator). The antenna is fed with RG-8A/U coax.

No balun is used nor was one tried. The antenna readily accepts the full output of my SB-200 linear amplifier for all swr readings of less than 2.6 to 1, and has performed very competitively on the air.

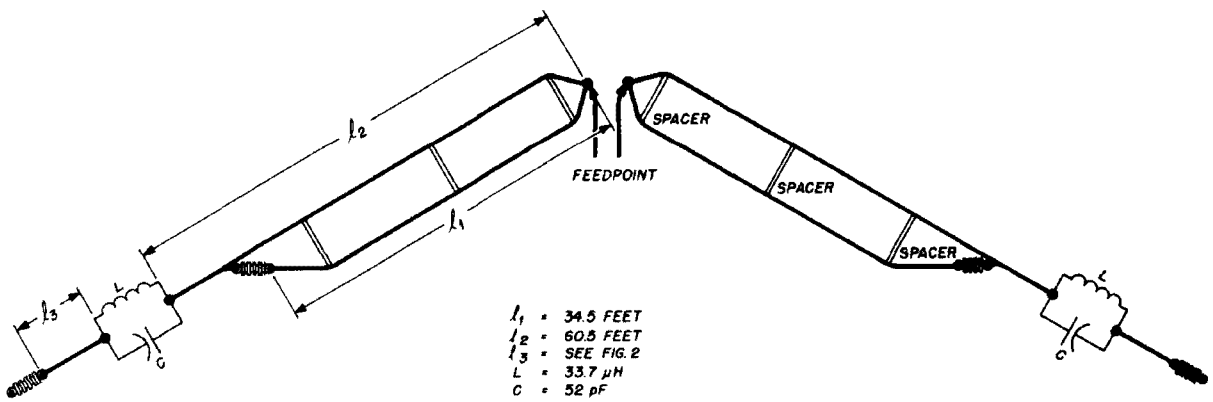


fig. 3. Converted-vee antenna system provides low swr operation on 80, 75 and 40 meters.

summary

This is the first time to my knowledge that parallel traps have been used to permit operation on two "in-band" frequencies. It is presumed that these results are applicable to horizontal antennas on 3.5 to 4.0 MHz although no data are on hand to illustrate this.

I wish to thank K6KA and W6EJJ for their encouragement, suggestions and assistance which contributed to the success of this venture.

reference

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ham radio

trapezoidal monitor scope for single sideband

Royden C. Golding, VE3CUS, 69 Gordon Road, Willowdale, Ontario, Canada

One of the advantages of single sideband is a reduction in interference because channel width is reduced by one-half. However, there seems to be more interference these days than when a-m was popular. Much of this can be attributed to increased activity. But I believe a large percentage of unnecessary interference from ssb equipment is the result of distortion the operator isn't aware of.

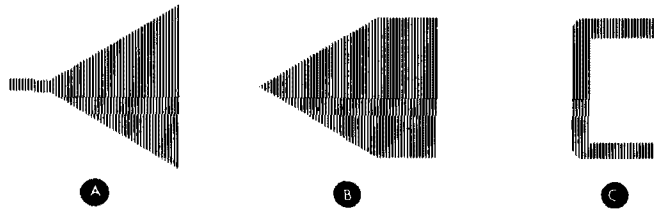
If a monitor scope is used in the conventional manner to display an ssb signal, the "Christmas tree" pattern will readily show distortion caused by **severe** flat-topping. However, **a small amount** of flat-topping, which may occur in the ssb exciter between the first mixer input and succeeding circuits, is virtually impossible to detect on the familiar "Christmas tree" scope pattern.

A good way to resolve this problem is to continuously monitor a trapezoidal pattern of the transmitter's performance. This article presents a simple circuit that can be used with a commercially manufactured monitor scope or with a home-built monitor using a small-diameter cathode-ray tube. When properly connected between your ssb transmitter and monitor, the circuit will allow a trapezoid pattern to be displayed. Such a pattern will respond to the slightest amount of flat-topping caused by nonlinearity in the transmitter.

An oscilloscope will show the wave envelope of an a-m or ssb transmitter if (a) a sample of the modulated signal is applied to the scope's vertical deflection plates, and (b) the scope's horizontal deflection plates are driven by a signal from a time-base generator. The time-base signal, or sweep, is usually generated inside the scope. If the scope doesn't have such a circuit, the 60-Hz line frequency can be used. The resulting scope pattern is useful for checking modulation percentage. It will respond to extreme nonlinearities, but it doesn't tell you much about what's happening in low-level stages. A trapezoidal pattern is necessary for this.

The trapezoidal pattern is actually a graphical representation of the transmitter's modulation characteristic. Any nonlinearity, however slight, becomes immediately apparent as the pattern shape departs from a true trapezoid. Typical patterns for a-m and ssb signals with over

fig. 1. Trapezoidal scope patterns. An a-m signal with over modulation is shown at A; ssb signal with flat-topping, B. C represents a saturated transistor or faulty diode in the monitor circuits.



modulation and flat-topping are shown in **figs. 1A** and **1B.** (The pattern in **fig. 1C** is caused by problems in the instrumentations circuit, which are discussed later.)

A trapezoid may be displayed on a scope by applying a sample of the transmitter's modulated output to the scope's vertical plates and applying a sample of the audio modulating signal to the horizontal plates. The signal applied to the scope's horizontal plates must have no rf

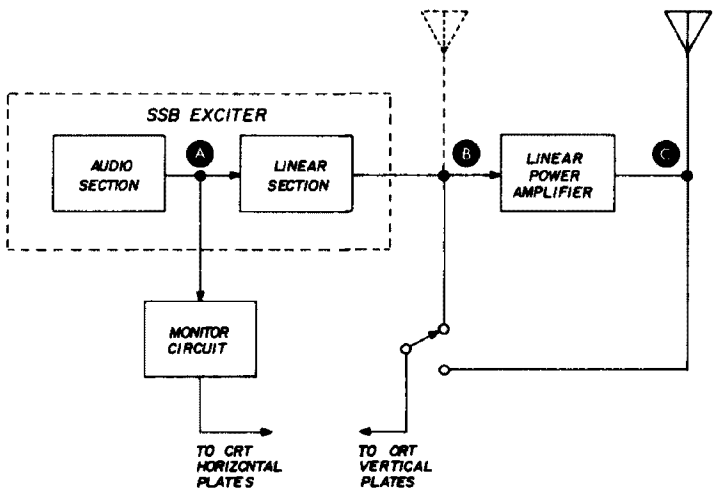


fig. 2. Block diagram showing low- and high-level inputs to oscilloscope. Low-level input, point A, is taken from grid tap on first mixer. High-level input may be taken from exciter output, point B, or from output of linear amplifier, point C.

component. Therefore, the modulating signal from your ssb exciter must be rectified before it's applied to the scope. This is the basis of the monitor circuit to be described. This circuit consists of a simple rf amplifier followed by a diode detector that drives an audio amplifier to provide sufficient voltage for the monitor scope's horizontal plates.

pick-off points

The place where the monitor circuit is inserted into the ssb transmitter is important. The block diagram of **fig. 2** will help clarify this. The input signal to the monitor circuit is taken from the exciter's **first mixer.** In most ssb exciters, this is the first stage that could produce nonlinear operation, and we wish to monitor **all** linear stages from this point on.

The second pick-off point, where the transmitter's modulated output is applied to the scope's vertical plates, can be at the exciter's output or at the output of a linear final amplifier. If a switching circuit is used, as shown in **fig. 2,** you can monitor the entire system. However, the important thing to remember is that the pick-off point for the accessory circuit should be at the input to the **first mixer.**

distortion in the audio circuits

If there is distortion in your ssb signal caused by the speech amplifier or balanced modulator, a more sophisticated circuit than that described here will be required to isolate the problem. However, if such distortion should occur, the result at the receiving end of your transmission is garbled speech, **not** spurious radiation caused by flat-topping. Splatter, caused by nonlinearities in low-level rf circuits, is what we're trying to monitor with a trapezoidal pattern on the scope.

preliminary checks

The first thing to do is to check your exciter and decide just how you're going to adapt the circuit. The input to the vertical plates of the scope is no real problem. It can be coupled from the exciter (or linear amplifier) output through a capacitor, using a convenient length of coaxial cable. Use a small link at the transmitter. The link can be adjusted to give an appropriate amount of signal.

The input to the scope's horizontal plates is applied through the circuit of **fig. 3.** Bear in mind that the output signal from this circuit has the same wave shape (less the rf component) as the signal applied to the scope's vertical plates.

Before starting construction, check your exciter to determine if you have enough spare drive for the monitor circuit. The signal should be picked off as soon as possible after the unwanted sideband has been eliminated. In some exciters this is the "first mixer" or "first i-f amplifier," or possibly the "balanced mixer." Only by

monitoring the signal from the grid of this tube to the output can we be sure of including all linear circuits.

Now decide where to mount the indicating scope tube (peeking out through a hole in the exciter front panel is an ideal arrangement). Measure the distance (this can often be reduced to four

construction

The rf transformer should be tackled first (**fig. 3**). The design will depend upon the carrier frequency of your rig. The transformer needs a bandwidth of perhaps 3000 Hz above and below the carrier frequency. If your carrier is 455 kHz, a suitable i-f transformer can be found in

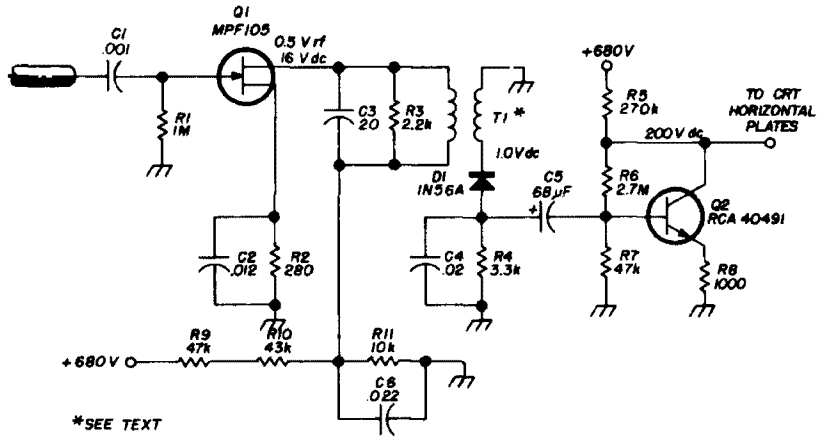


fig. 3. Simple three-stage monitor circuit for driving horizontal plates of scope to obtain trapezoidal pattern. Power is taken from exciter supply.

or five inches) from the grid of the first mixer to the nearest point in the set where you can mount the circuit board. Cut a piece of coax to this length, then ground the grid of the first mixer through the coax, a .01 µF ceramic capacitor, and a 1-megohm resistor (**fig. 4**).

Now try the set under operating conditions. If the mixer grid is tuned, you will have upset the tuning. Retune the circuit until full output is again obtained. This proves that the set will take the slight drain needed for the monitor circuit and still operate. If the mixer grid is not tuned, probably no effect will be noticed. But if output is reduced, some means of increasing drive at the mixer grid must be found to compensate for the loss. If it is a balanced mixer circuit, try placing equal capacitance and resistance on the second grid to keep the two halves balanced.

the surplus shops. A 9-MHz carrier is also fairly common. A successful 8-MHz mc coil was made by adding a few turns to the primary of a 9-MHz i-f transformer until it resonated at 8-MHz with a 20-pF ceramic capacitor and a 2200-ohm resistor across it. The secondary coil should have more inductance than the primary coil to give a voltage step-up. A gain of 1½ to 2 is adequate.

We now have some inductance in the circuit. This is desirable, since capacitance introduces phase shift and consequent distortion in the scope pattern. At all points we must watch linearity in the design. Our whole purpose is to demonstrate linearity on the screen; thus any distortion introduced in the monitor circuit would give a false indication.

An input rf stage is needed as well as an audio amplifier to follow the demodulating diode.

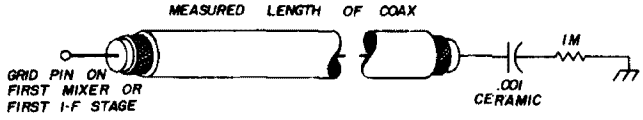


fig. 4. Test circuit to determine drive capability of exciter. It's used once and discarded.

We need a high impedance input for the MPF105 jfet which is provided by R1. For the output we need a high resistance; this is provided by the tuned primary transformer circuit. The diode is quite critical, particularly as to temperature. Out of a large number tried, the one selected was a 1N56A.

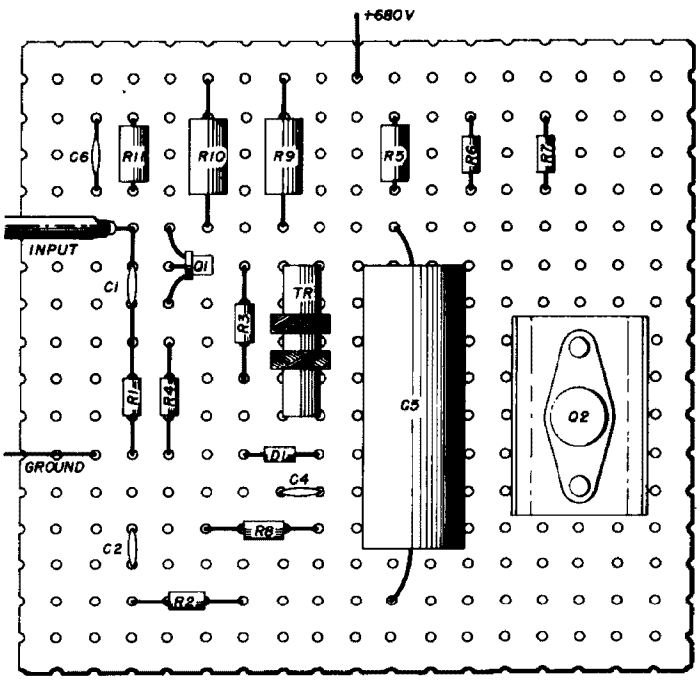


fig. 5. Arrangement of components for the monitor.

The coupling capacitor feeding the signal to the power transistor must be large to keep down reactance and phase shift.

The output transistor operates at a rather high voltage.* Audio output under the conditions described should be too high rather than too low. Rf voltages at certain parts of the circuit are given in **fig. 3.** These were measured with a Heathkit vtvm. A series of "aah's," spoken close to the microphone, will produce the values shown.

* The author used an RCA 40264 transistor for the audio amplifier (also shown in the circuit-board photo). This device is no longer produced by RCA. However, the RCA type 40491 is an acceptable substitute. Its electrical characteristics are almost identical to those of the 40264. Also, the 40491 has a heat radiator, whereas the 40264 does not. The 40264 will fit nicely on the circuit board. It costs about 85c and is available from Allied Radio. **Editor.**

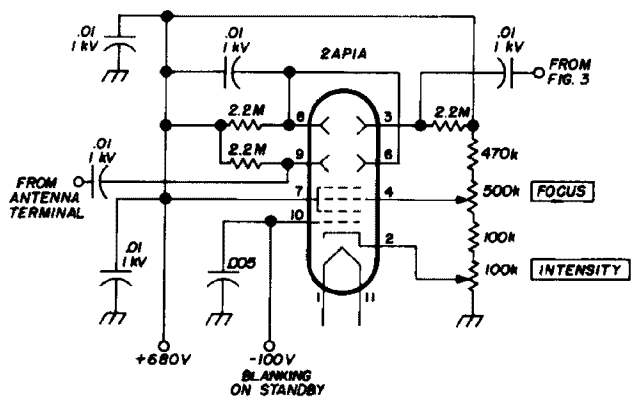
If too much voltage is obtained, decrease R4's value until a suitable pattern size is obtained. An unusual scope pattern obtained at one time is shown in **fig. 1C;** this was the result of a faulty diode.

A suggested component layout for the monitor circuit is shown in **fig. 5.** The Vector board may be mounted inside your exciter, as discussed previously. All signal leads should be coax cable (such as RG-58/U), and the shield should be grounded at each end.

My original scope circuit is shown in **fig. 6.** I've used this circuit, so I'm certain it will work. The scope circuit shown in the 1969 edition of the ARRL Handbook, however, should be equally satisfactory. Note that the voltages in **fig. 6** are lower than those shown in the Handbook circuit. These voltages were available in my transmitter power supply, so I used them.

In **fig. 6,** the 2AP1A grid is biased to blank the spot when the transmitter is on standby. You should be able to pick up —100 volts in your ssb exciter for this purpose.

fig. 6. Monitor scope circuit using transmitter high-voltage supply. Any good scope circuit could be used.



If you wish to use a commercially manufactured monitor, the amplifiers will have to be disconnected and the signals fed directly to the scope plates.

I think you'll find the circuits described here will provide a good indicator to show when your rig is flat-topping. I encourage you to use it all the time.

ham radio

graphical solution of impedance-matching problems

Using simple geometry, these methods allow you to design a variety of networks

I. L. McNally, W1NCK, 11 Jackson Street, Marshfield, Massachusetts 02050
Henry S. Keen, W2CTK, 64 Schuyler Drive, Commack, New York 11725

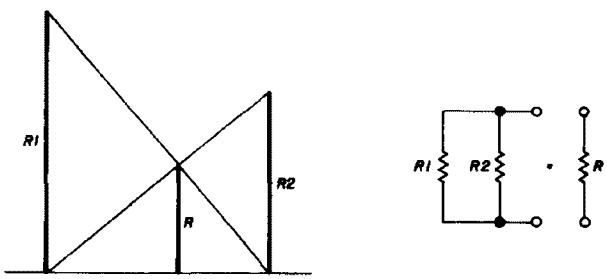
One of the most common problems in radio circuits is matching one impedance to another. The problem might be that of matching a transmitter output stage to a resistive load, or the load may have a reactive component, as is usually the case when attempting to transfer power to an antenna.

Many articles have been written covering the mathematics of this problem and also the application of the Smith chart.[1,2,3] Impedance-matching problems can be solved readily with sufficient accuracy for practical purposes with no more equipment than a straight edge, compass and graph paper. The graphical method lends itself to multiple-component networks involving complex impedances, without resorting to trigonometry or complex algebra. It allows a visual choice of constants and shows forbidden approaches in choosing impedance paths.

The method presented in this article will allow you to solve most impedance problems encountered in amateur work. The geometric principles are easy to follow, and you'll need to make only a few simple computations. Rules are given for constructing the diagrams. Typical examples and solutions are shown. The examples are presented without mathematical proof, however. For those who wish to pursue the classical approach, some excellent material will be found in references 4, 5 and 6.

a starting point

First consider the familiar methods known as the "leaning ladder" diagram for determining the resultant of two resistors or reactances in parallel (**fig. 1**). Two perpendicular lines with lengths proportional to the two resistors or reactances are erected with arbitrary separation from a common baseline. Lines are then drawn from the top of each perpendicular to the base of the other. A third perpendicular is now drawn from the intersection of these lines to the baseline. The length of this new perpendicular is proportional to the combined resistance or reactance of the two parallel elements.

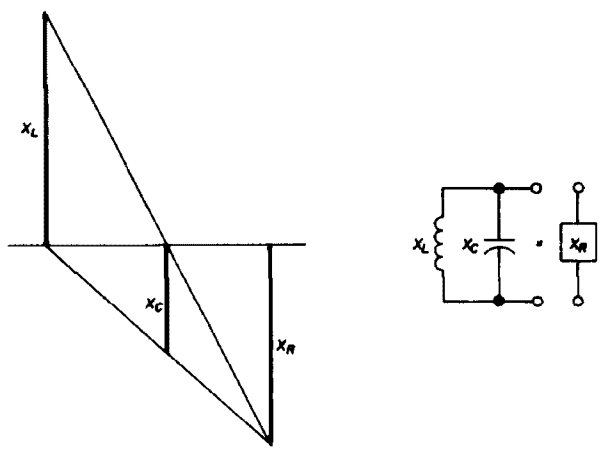


1. The "leaning ladder" diagram for finding the resultant of two resistences or reactances in parallel.

What happens, however, when two reactances of opposite sign are to be evaluated? The same procedure is followed as before, except that the perpendicular lines representing the reactances will be located on opposite sides of the baseline (**fig. 2**). Again we connect the end of each perpendicular to the base of the other, extending the lines until they intersect. The length of a perpendicular from this point of intersection to the baseline represents the combined reactance of the two paralleled elements. The side of the baseline where the intersection takes place determines whether the resultant, X_R, is inductive or capacitive.

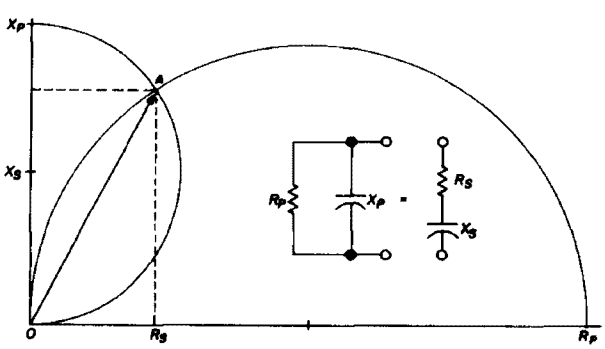
Now suppose a reactance is to be paralleled with a resistance. How do you determine the impedance of such a combination? Semicircles are constructed upon rectangular coordinates, with diameters proportional to the paralleled resistance and reactance, intersecting at point A (**fig. 3**). A line, O-A, from the origin to the point of intersection will be proportional to the impedance of the combination. The projections of this point of intersection upon the resistive and reactive axes will then be proportional to the resistance, R_S, and reactance, X_S, respectively, which make up the series equivalent of the parallel combination.

Because an angle inscribed in a semicircle is always a right angle, it is easily shown that the point of intersection, A, lies on a straight line connecting the ends of the two diameters. This construction leads to a well-known diagram frequently used to solve L networks, (**fig. 4**). An L network is merely a transformation from a parallel resonant circuit, seen looking in at Z1, to a series resonant circuit, seen looking in at Z2.



2. Diagram for finding resultant impedance of two reactances of opposite sign.

3. Geometry for solving parallel-to-series transformion.



rules for construction

By combining these diagrams, it's possible to solve a variety of matching-network problems. The geometry of **fig. 5** is the basis of solving all problems using this method. General rules for using the method are:

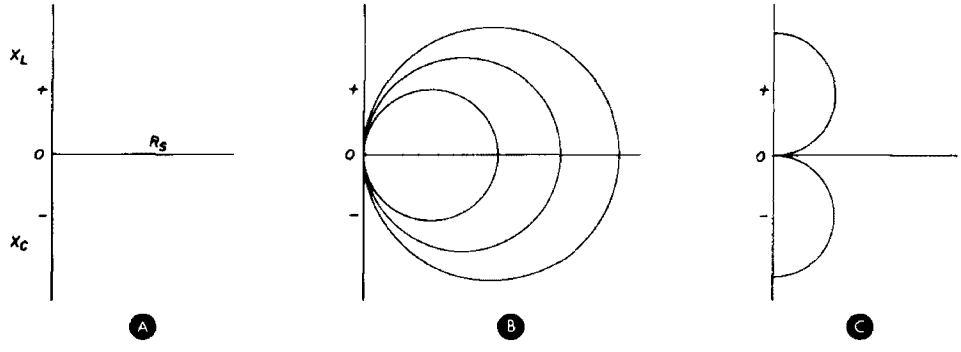
1. Adding a series of reactances moves the impedance on a vertical line: up for inductive and down for capacitive reactance.

2. Adding a parallel reactance moves the impedance along a circle with its center on the horizontal axis. It rotates clockwise for capacitive and counter clockwise for inductive reactance.

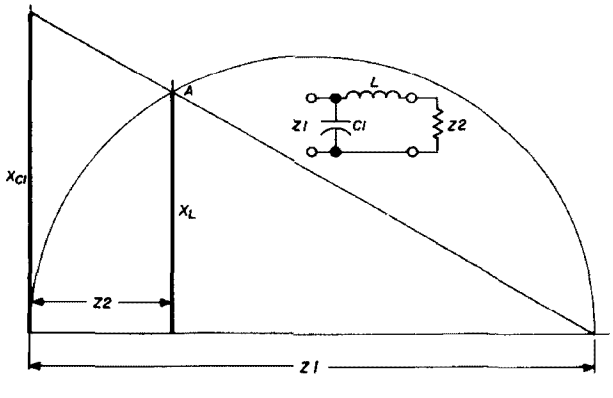
3. When choosing impedance paths, it is not permissible to use a path passing through the origin of coordinates.

The method permits rapid comparison of different network designs without a knowledge of complex algebra, and a clear picture is given of what happens when parameters are modified.

We'll begin with the pi network since this is one of the most-used circuits in amateur work. Other circuits will then be described, which will provide a foundation for solving most impedance-matching problems. Some numerical examples are then given to show step-by-step procedures.



5. Basic geometry for graphical solution of impedance problems. Series resistance is plotted along the horizontal axis and series reactance on the vertical axis, as at (A). Parallel resistance and parallel reactance circles are constructed as in (B) and (C) respectively.



4. The L network diagram. Another example of parallel-to-series transformation.

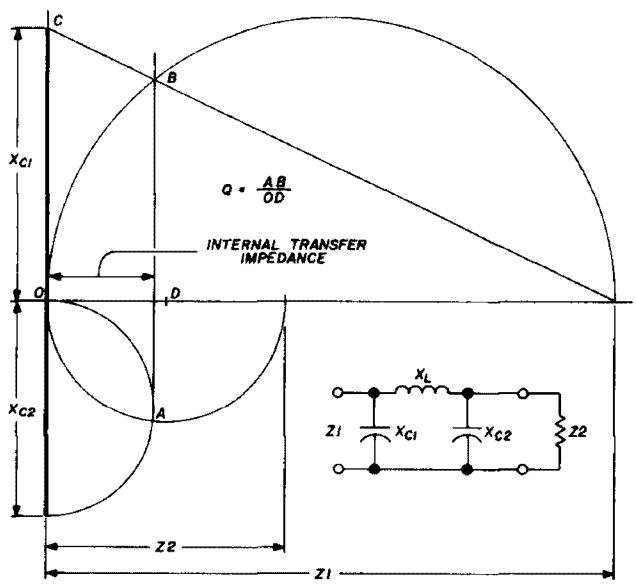
the pi network

The pi network can be considered as two cascaded L networks, designed to transform both input and output impedances to a common internal transfer impedance, which must be lower than either terminal impedance. This internal transfer impedance determines the network Q, a fact that becomes apparent from a consideration of the design diagram (**fig. 6**).

To design a pi network, begin at the origin of a set of rectangular coordinates, and construct a semicircle above the horizontal axis, with diameter proportional to Z1. (Z1 is the greater of the two terminal impedances.) Similarly, from the origin con-

struct a second semicircle below the axis. Its diameter is proportional to Z2, the lesser of the two terminal impedances.

Because an infinite number of solutions exist to a pi-network problem when terminal impedances are specified, an assumption must be made for one of the three reactances. This is necessary to establish the internal transfer impedance. There are certain advantages if the reactance of the output capacitor, C2, is made equal to the load resistance, Z2. However, network Q requirements frequently dictate a lower value as discussed later.



6. The pi network diagram. Circuit Q is determined by the internal transfer impedance.

The assumed reactance, X_{C2}, of the output capacitor becomes the diameter of a third semicircle, beginning at the origin and constructed downward below the horizontal axis. The point of intersection between this and the Z2 semicircle is point A. From this point a vertical line is drawn to intersect the original Z1 semicircle at point B. The length of the line segment, AB, represents the required reactance of X_L.

A straight line is now drawn from the extreme end of the Z1 diameter through point B, intersecting the vertical axis at point C. Line OC will then be proportional to X_{C1}, the reactance of the required input capacitor.

The intersection of inductive reactance line AB with the horizontal axis is point D. The significance of this point is that line segment OD represents the internal transfer impedance of the network. The Q of the network, when driven by a current generator such as a screen-grid tube or a transistor, will be equal to the inductive reactance, AB, divided by the internal transfer impedance, OD. The Q will also be equal to $Z1/X_{C1}$ plus $Z2/X_{C2}$, which can be proven identical.

When driven by a resistive source, such as a triode, the network is loaded from both ends, and the effective Q may be cut in half.

the tee network

Although the T network is not as well known as the pi network, it is a very useful circuit and is quickly solved graphically. With the T network, we may assume the internal transfer impedance as equal to or greater than the sum of the terminal impedances Z1 and Z2, usually by a factor or two or more. The graphical design procedure, with reference to **fig. 7** is as follows:

Construct a semicircle with horizontal diameter **greater** than the sum of the terminal impedances. Mark off, from opposite ends of the diameter, line segments proportional to the two terminal impedances. From these two points erect perpendiculars to intersect the semicircle at points A and B respectively. Connect points A and B to the remote ends of the diameter, intersecting each other at C. A perpendicular, CD, to the diameter will be proportional to the reactance of the capacitor, C. The Q of this network, as driven by a current generator, will be equal to the diameter of the semi-circle divided by the line segment CD. The sum of $X_{L1}/Z1$ plus $X_{L2}/Z2$ will give an identical result.

Although the derivation of this diagram may seem obscure, if perpendiculars are erected at the ends of the diameter, and the slant lines extended to intersect these perpendiculars, we will have the two superimposed L-network diagrams. The line segments of these end perpendiculars will each represent a capacitive reactance correspond-

ing to one of the two cascaded L networks making up the complete T network. The extended slant lines can then be seen to represent the leaning-ladder diagram, with line CD being the result of both capacitive components in parallel. Therefore all construction exterior to the semicircle will be redundant and can be omitted.

If the semicircle is constructed so that its diameter is equal to the sum of the two terminal impedances, all reactances will be of the same magnitude, differing only in sign, and will be equal to the geometric mean of the terminal impedances. The Q of such a network would be quite low, being equal to the sum of the two terminal impedances, divided by the square root of their product.

When the pi network is designed so that $Z2 = X_{C2}$, excursions of Z2 will have minimum effect on Z1. Resonance will be maintained by retuning X_L. A network is possible whereby a two-to-one range of Z2 (assumed purely resistive) will, in turn, cause Z1 to vary from the target impedance by less than five percent.

Similarly, design of the T network so that $Z1 = X_{L1}$ will permit Z1 variations of the same magnitude, with the network output still presenting a match to the load (within the same limits). Resonance is maintained by retuning C1.

In either network the terminal impedance, Z1, is assumed the higher of the two. Although Z1 has been treated at the input end, either network is completely reciprocal.

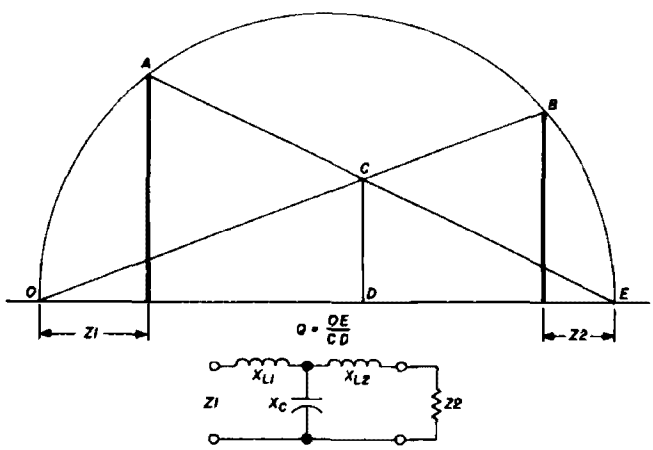
examples using pi networks

problem 1. Match 3000+j0 ohms to 600+j0 ohms with a pi network. In this case, Q = 10.

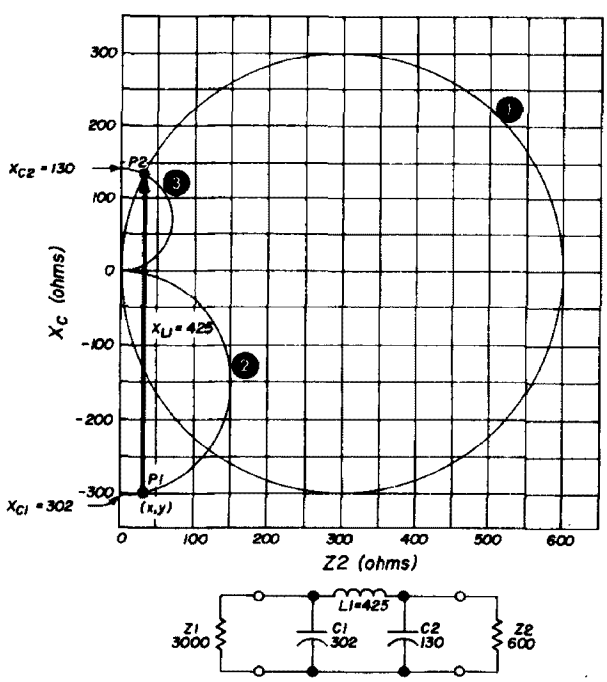
1. Draw a 600-ohm circle (1), ***fig. 8.***
2. Calculate point (x, y) and plot:

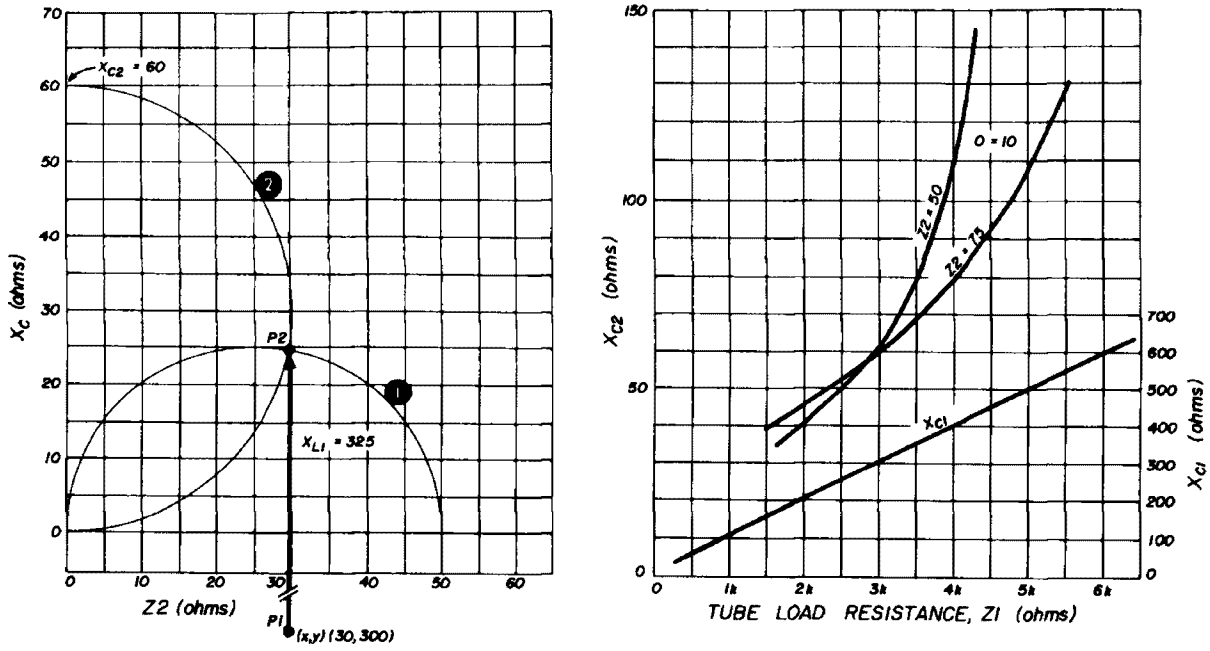
$$y = \frac{R1}{Q} = \frac{3000}{10} = 300.$$

7. Solving the T network. The internal transfer impedance is equal to or greater than Z1 + Z2 by a factor of two.



8. Solution to problem 1: matching 3000+j0 ohms to 600+j0 ohms with a pi network. Arrows indicate impedance path.





9. Solution to problem 2: matching 3000+j0 ohms to 50+j0 ohms with a pi network——a common problam in transferring tube output impedance to an antenna transmission line. Expanded scales for R2=50 ohms are shown at (A). At (B) the curves are limited to R2=50 or 75 ohms and Q=10.

$$x = \frac{R1}{Q^2} = \frac{3000}{100} = 30.$$

3. Erect a vertical line from point P1 (x, y) to intersect the 600-ohm circle at point P2. This is X_L to scale (425 ohms).

4. Draw circle (2). Its intercept on the vertical axis at 302 ohms is X_{C1}.

5. Draw circle (3). Its intercept on the vertical axis at 130 ohms is X_{C2}.

6. Solution: L1 = 425 ohms
   C1 = 302 ohms
   C2 = 130 ohms

problem 2. This is the same as problem 1, except Z2 = 50 ohms (**fig.** 9).

1. Draw a 50-ohm circle (1), **fig. 9A**.

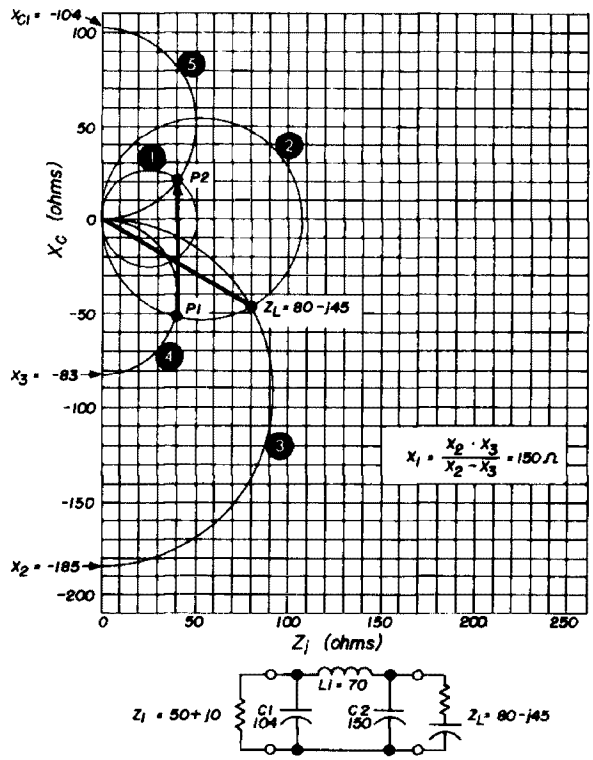
2. Erect X_{L1} through X = 30 to intersect the 50-ohm circle at point P2. This scales to y + 25, or X_{L1} = 325 ohms.

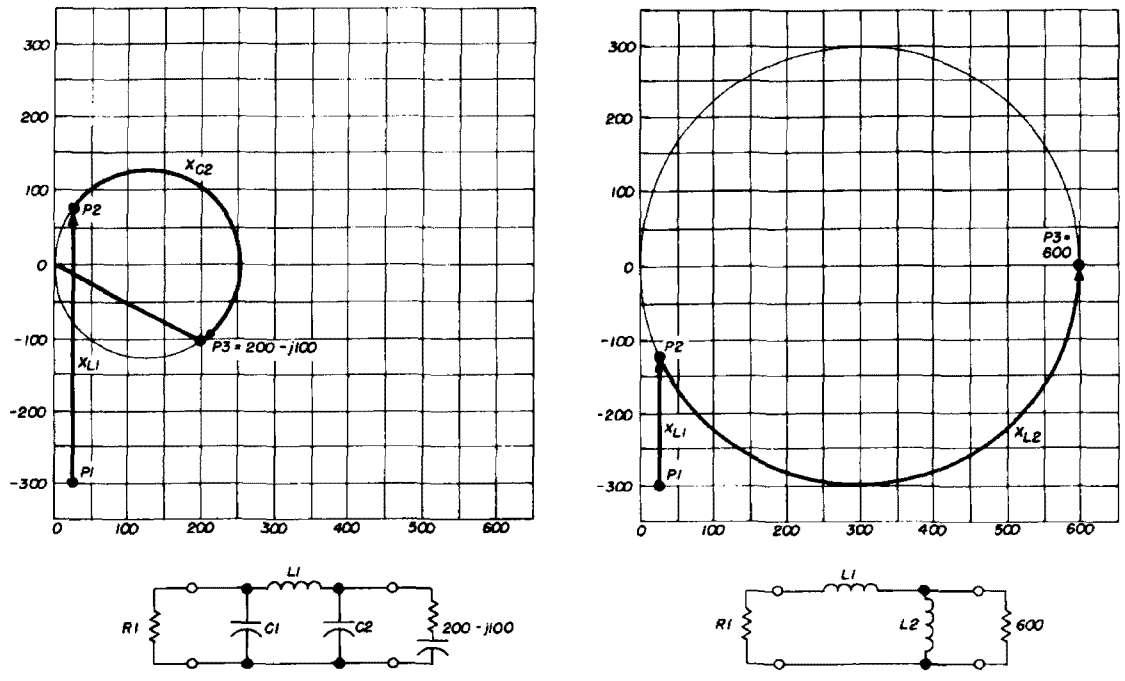
3. Draw circle (2) through point P2. It will intersect the vertical axis at X_{C2} = 60 ohms.

4. Solution: L1 = 325 ohms
   C1 = 302 ohms
   C2 = 60 ohms

Some pi-network curves for common tube load resistance are shown in **fig. 9B**.

10. Solution to problem 3: matching an antenna load of 80—j45 ohms with a pi network. This has a three step impedance path: ZL to P1 to P2 to Zi.





11. Examples of correct impedance paths. Both show networks for matching typical power amplifier tube impedances to various loads. Note that the impedance paths must not pass through the origin of coordinates.

problem 3. Match an antenna load of 80-j45 ohms to 50 ohms using a pi network (**fig. 10**).

1. Construct a 50-ohm circle (1).
2. Plot $Z_L = 80\text{-}j45$ ohms.
3. Construct circle (2) through Z_L.
4. By inspection the maximum value of X_L is about 75 ohms. Select a value of 70 ohms and fit it vertically as P1, P2 between circles (1) and (2).
5. Construct circle (3) through Z_L. It will intersect the vertical axis at —185 ohms (X2).
6. Construct circle (4) through P1. It will intersect the vertical axis at —83 ohms (X3).

$$X1 = \frac{X2 \cdot X3}{X2 - X3}$$

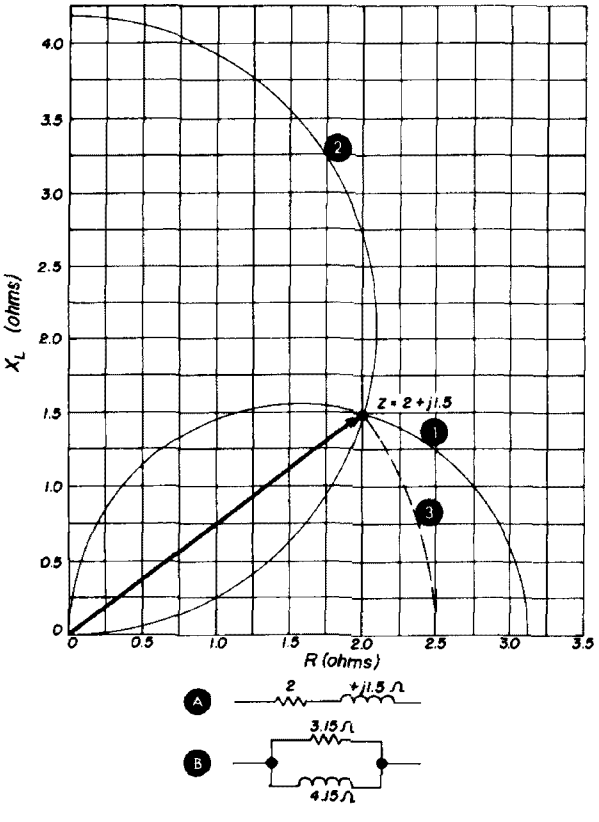
$$= \frac{(-185)\ (-83)}{(-185) - (-83)} = \frac{-185 \cdot 83}{102}$$

$$= 150 \text{ ohms.}$$

This is capacitive reactance added by moving from Z_L to P1.

choice of impedance paths

Recall that the internal transfer impedance must be lower than either terminal im-



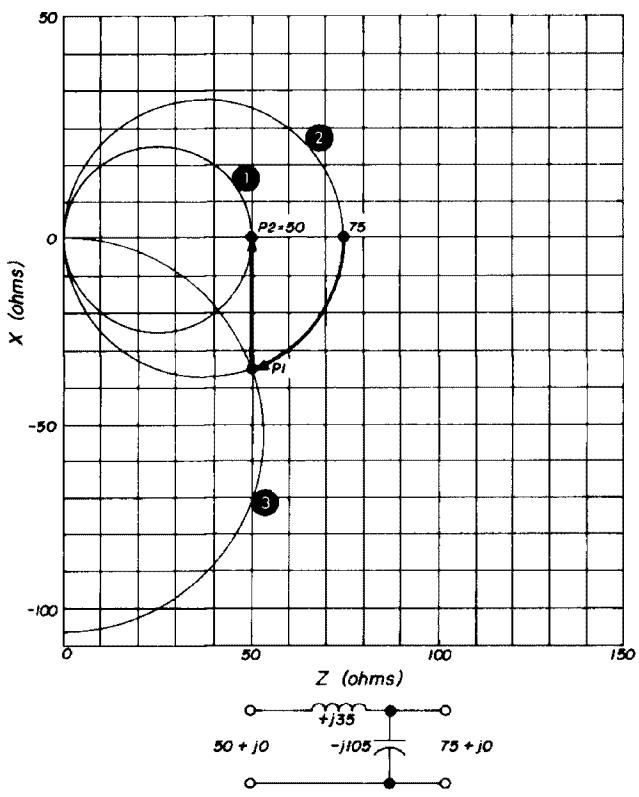
12. Solution of series-to-parallel transformation. The resultant impedance is 2.5 ohms.

pedance. The internal transfer impedance determines the Q of the pi network. From **fig. 6,** line segment OD determines this parameter. Therefore, from the rules of construction for graphical solution to these problems, it is not permissible to choose an impedance path through the origin of coordinates. Examples of correct impedance paths are shown in **fig. 11.** Here, these show transformation between typical tube output impedances and various load impedances. Note that the paths do not pass through the origin. While a 600-ohm terminal impedance is not too common in most rf circuits these days, the example does indicate the principles to be followed when designing these networks.

solving series and L networks

problem 1. Given the series circuit of **fig. 12A,** find the equivalent parallel circuit.

1. Plot the impedance vector Z = 2 + j1.5.
2. Construct circle (1) with its center on the horizontal axis, which passes through the origin and Z as shown. It will intersect the horizontal axis at 3.15 ohms resistance.
3. Construct circle (2) with its center on the vertical axis, which passes through the origin and Z. It will intersect the vertical axis at 4.15 ohms inductive reactance. The equivalent parallel circuit is shown in **fig. 12B.**
4. Solution. Scaling the Z vector gives an impedance of 2.5 ohms (3).



13. Examples of 2-step impedance path from 75 ohms to 50 ohms in an L network.

problem 2. In the network of **fig. 13,** it is desired to find X_L, X_C, L, and C for a frequency of 3.9 MHz.

1. Construct circles (1) and (2) through 50 ohms and 75 ohms as shown.
2. Construct line P1-P2, which is X_L series and scales 35 ohms.
3. Construct circle (3) through P1. It will intersect the vertical axis at —105 ohms. This is X_C parallel capacitive reactance obtained in moving clockwise from 75 ohms along circle (2) to P1, which is directly below P2, the 50-ohm point.
4. Solution.

$$L = \frac{X_L}{2\pi f} = \frac{35}{2\pi \times 3.9} = 1.43\ \mu H$$

$$C = \frac{10^6}{2\pi f x_C} = \frac{10^6}{2\pi \times 3.9 \times 105} = 386\ pF$$

$X_L = 35$ ohms

$X_C = 105$ ohms

problem 3. Match 50 + j0 ohm to Z_L = 125-j72 ohms, a complex load (**fig. 14**).

1. Construct a 50-ohm circle (1).
2. Plot $Z_L = 125 - j72$.
3. Construct circle (2) through Z_L.
4. Construct a vertical line from P_2 to P_1. This is the series X_L and scales 77 ohms.
5. Construct circle (3) through the origin and Z_L. It will intersect the vertical axis at —285 ohms (X2).

6. Construct circle (4) through the origin and P2. It will intersect the vertical axis at —110 ohms (X3).

7. When neither of the terminal points is on the horizontal axis, as in this case with Z_L and P2, it is necessary to compute the value of reactance involved in moving from Z2 to P2.

$$X1 = \frac{X2 \cdot X3}{X2 - X3}$$

X1 = added reactance

X2 = initial reactance

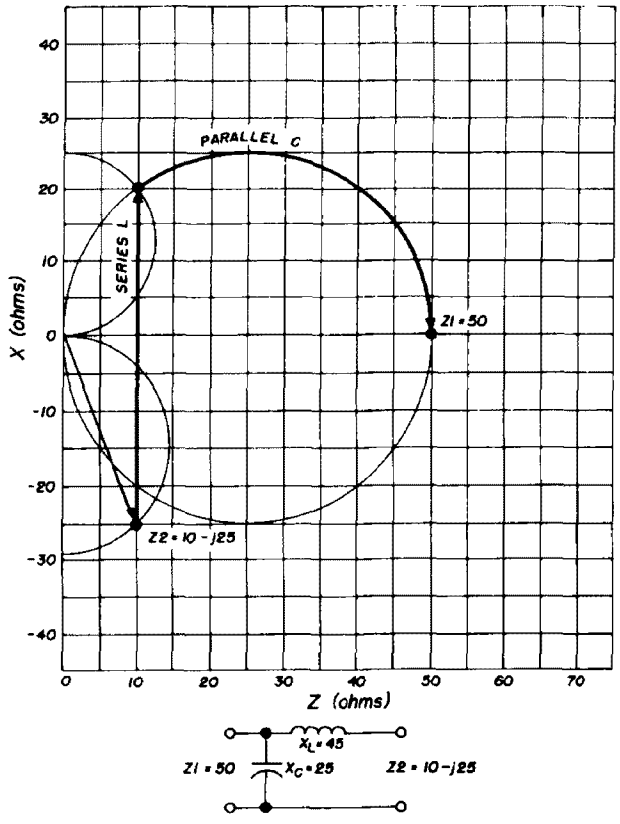
X3 = final reactance

$$X1 = \frac{(-285)\ (-110)}{(-285) - (-110)}$$

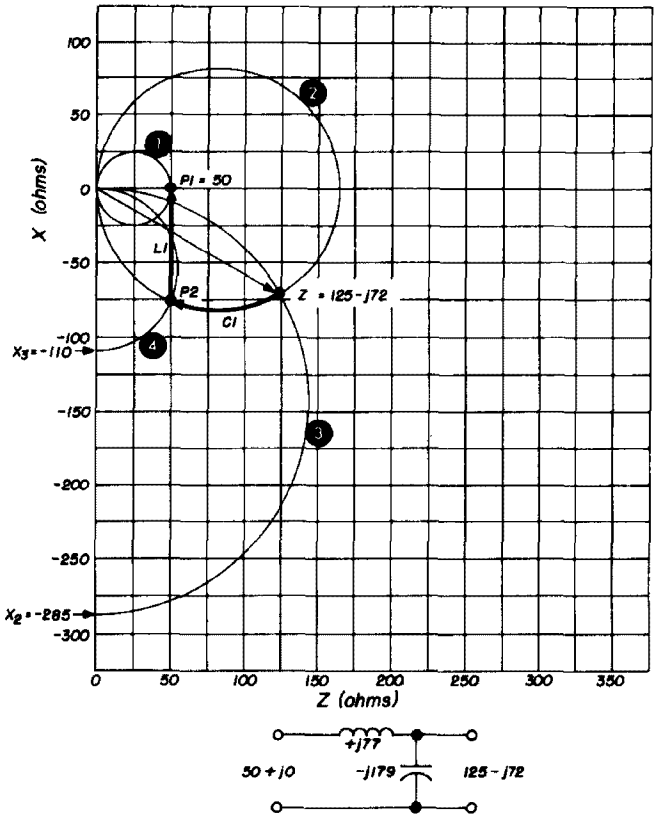
$$= -179 \text{ ohm capacitive}$$

8. Solution: L1 = j77 ohms, and C1 = j170 ohms.

In **fig. 15** are examples of three different



15A. Impedance path through matching network consisting of series L and parallel C.



14. Matching a 50-ohm resistive load to a complex load. The impedance path is from ZL to P2 to P1.

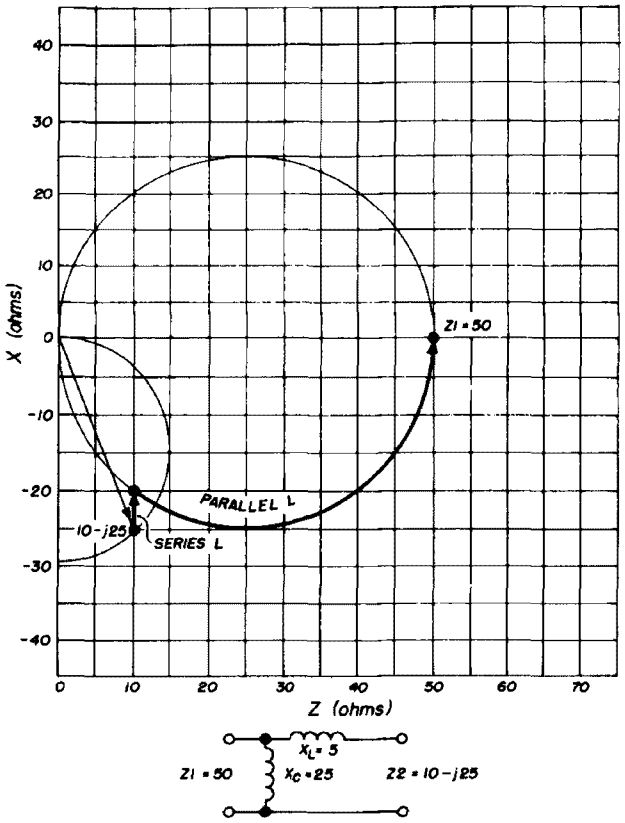
choices of impedance paths in going from Z2 to Z1 using three network combinations to match a resistive 50-ohm load to a complex load of 10-j25 ohms. Again, the "forbidden" path is not to be used because it passes through the origin.

summary

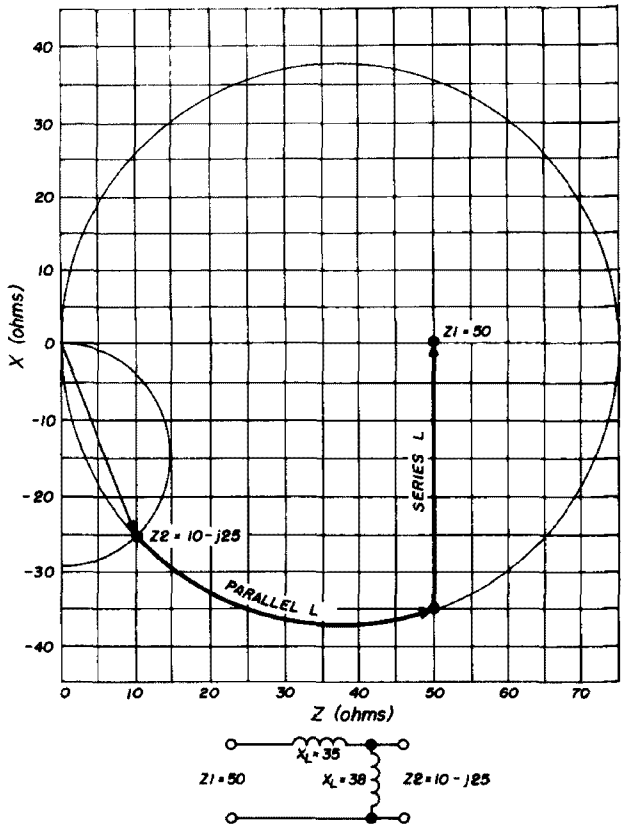
We have shown examples of solving the most common impedance-matching problems using simple geometric methods. The following notes are offered in adopting these methods for solving a wide variety of problems.

1. The L-network is one of the most useful circuits known for matching nearly all direct-coupled tank systems. The examples show how to match a high-resistance to a low-impedance reactive load. If the converse is desired, it is only necessary to convert the reactive load to its equivalent parallel components.

2. T networks are useful as harmonic attenuators in low-impedance transmission



15B. Impedance path through matching network consisting of series and parallel L, respectively.



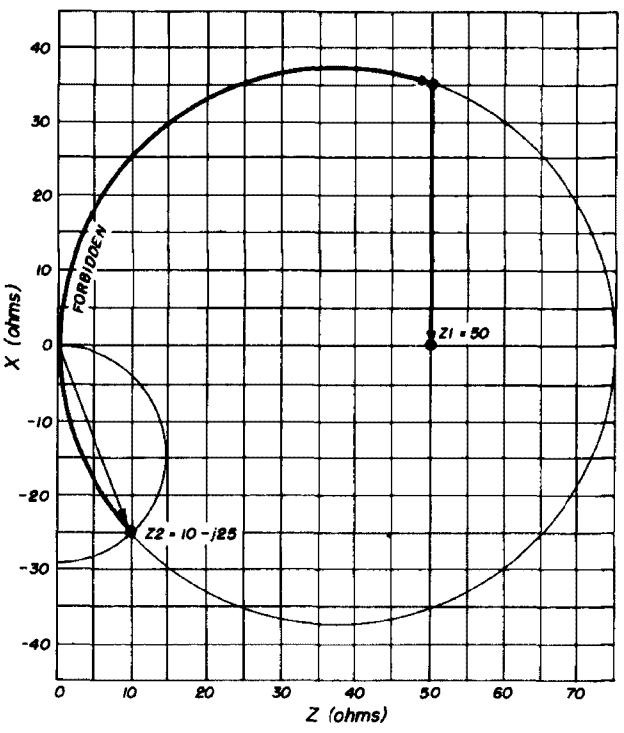
15C. Impedance path through parallel and series L, respectively.

lines. These can be readily solved by treating them as two cascaded L sections and combining the capacitances.

3. The pi network is used to match a wide range of load impedances with reasonable tank-circuit Q. Contrary to some popular notions, the pi network will not match a tube to any length of wire. The circuit is load-limited by the ratio of tube load and circuit Q if it is to perform as an efficient transformer.[5]

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15. Another example of impedance paths in network design. Three different 2-step paths are shown from Z2 to Z1 as well as the "forbidden" path through the origin.

new life for the Collins 51J receiver vfo

The local oscillator in this excellent receiver can be restored to like-new operation with these modifications

William I. Orr, W6SAI, Eimac Division of Varian, San Carlos, California 94070

The Collins 51J series receiver (military type R-388/URR) is a valuable adjunct to any amateur station. Several articles have described modifications to make this superb receiver better suited to single sideband reception.[1,2,3]

Available through surplus sales or MARS issue, the 51J receiver offers low-drift frequency readout capability from 500 kHz to 30.5 MHz. Indeed, a mint condition 51J with its 100 kHz crystal calibrator makes a good frequency meter for the home laboratory or amateur station in addition to providing excellent high-frequency and broadcast-band reception.

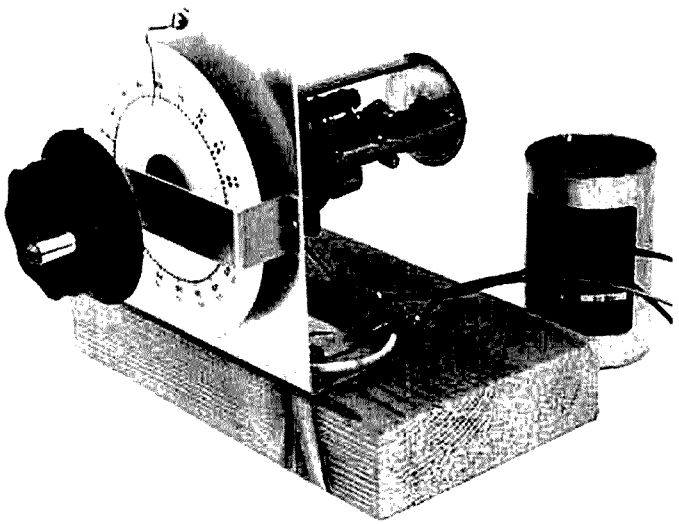
I've noticed that secondhand 51J receivers seem to be valued in proportion to the accuracy of tuning-dial calibration which, in turn, is a direct function of tuning linearity of the variable frequency oscillator (vfo)—the heart of the receiver. The better the vfo tracking characteristics, the greater the worth of the receiver. The difference in cost between a receiver having a good and a poor tracking characteristic is several hundred dollars.

vfo calibration shift

My 51J receiver was purchased secondhand in 1956. At that time, I made an accurate frequency check of the vfo and filed

the results. Over the years, I noticed a gradual, slow shift in calibration linearity. Finally, in early 1968, I found that correction was beyond the capability of the dial zero adjust (or fiduciary setting, as it's sometimes called).

A check with other owners of older 51J's indicated that their receivers had the same slow drift to a certain extent. Interestingly enough, in all cases, the calibration shift was in the same direction and of the same magnitude. That is, the vfo tuning range



Jig for Collins vfo adjustment. Oscillator cover is removed to expose adjustment coil, L-002. Shield must be replaced for calibration, as it shifts resonant-circuit frequency.

gradually contracted. Instead of tuning 1000 kHz with ten revolutions of the tuning dial, the tuning range was reduced to about 990 kHz with ten revolutions. This made each end of the dial in error by 5 kHz when calibration was established at the center of the tuning range. For example, when calibration was "on the nose" at 15 MHz, at 14.5 MHz the dial would read 14.505 MHz; and at 15.5 MHz, it would read 15.495 MHz.

The 51J instruction manual is full of exhortations about not tampering with the sealed vfo, otherwise frequency calibration will be degraded. True, there's an auxiliary adjustment on the front of the vfo that can be used to restore tracking. However, in most older 51J's, the adjustment will have been "used up," and no further adjustment is possible.

modifications

This being the case, I felt my receiver's calibration was so far off that no further damage could be done to the sealed vfo unit if it was carefully opened and examined. Accordingly, the vfo was removed from the 51J (not without some trepidation), opened, modified, resealed, and replaced in the receiver. Results were so gratifying that I prepared this article to describe the modification, with the hope that it will benefit other owners of this fine receiver.

It should be noted that Collins approval of this radical surgery has not been obtained, and I doubt if such a modification, as described here, would meet with other than raised eyebrows. However, many 51J receivers have outlived their warranty. So if yours is one of these and suffers from the ailment described, the modification will restore receiver dial calibration to that comparable with a new set.

While modifying the vfo, several capacitors that have a notoriously short life can be replaced, thus enhancing the receiver's long-term reliability.

the variable frequency oscillator

This modification involves the Collins 70E-15 oscillator used in the 51J receivers. The unit tunes 2 to 3 MHz, with a generous overlap at each end of the range. Shaft rotation requires ten turns to cover the 1000-kHz range. Output voltage into an open load varies between 3.5 and 6 volts, at an operating potential of 150. Oscillator tube V-001 (see fig. 1) receives in-phase feedback voltage from amplifier tube V-002's screen circuit through capacitor C-008. The oscillator tank consists of trimmer inductor L-002, permeability-tuned inductor L-001, and tank capacitor C-001 in parallel with temperature compensating capacitors. The oscillator is tuned by moving the powdered iron core within inductor L-001. The core, or slug, traverses a lead

screw rotated by the receiver tuning dial. Tuning ranges and other data for the Collins vfo's are listed in **table 1.**

The over-all oscillator tuning range is established by adjustment of trimmer inductor L-002, which may be manipulated through an adjustment hole in the front

points have shrunk, and neither trimmer coil L-002 nor the zero adjust compensation device on the tuning dial will bring the end points within the correct range. When new, the vfo may be adjusted over a range of about 40 kHz, centered about the 1000-kHz tuning range (**fig. 2**). As the unit ages, the

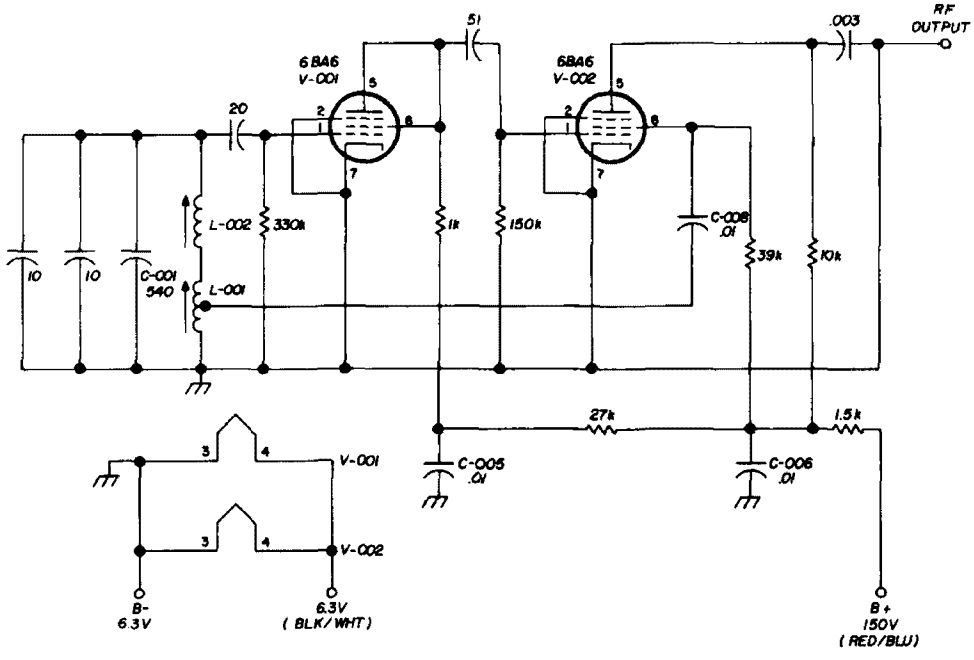


fig. 1. Collins 70E-15 vfo schematic. As tuning slug in L-001 ages, oscillator tuning range decreases until trimmer L-002 provides no more correction; L-002 is reworked to restore correct tuning range.

plate of the oscillator. With the aid of a special adjustment tool, L-002 may be tuned without removing the oscillator from the receiver. Vfo alignment by adjustment of coil L-002 is covered in the Collins instruction and maintenance manual (section 5.3.15).

When the receiver is new, this adjustment permits expansion and contraction of the vfo tuning range from a minimum of about 980 kHz to a maximum of about 1020 kHz. Proper adjustment of L-002's slug permits the tuning range end points to be moved sufficiently to ensure good calibration at the ends and center.

Alas, as the 51J receiver ages, it seems that the powdered-iron core of the main tuning slug ages also. At some future time, it will be found that the tuning range end

total range of L-002 remains about the same, but the entire range shifts in frequency as shown, until it doesn't include the desired correction range.

Luckily the slow shift of the vfo tuning range seems to decrease with time. Eventually stability is achieved, but at just about the time the user finds he has run out of vfo compensation because of the combined aging of the main tuning slug and the relatively restricted adjustment range of coil L-002.

An astute student of Murphy's Law could have predicted this impasse; thus the question is raised: what can be done to re-establish the correct tuning range of trimmer coil L-002 so that an older 51J may be brought into calibration and allowance provided for future corrections, if needed?

vfo removal

In order to work on the vfo, it's necessary to remove it from the receiver and mount it in a temporary calibration jig, as shown in the photograph. To begin this task, remove the vfo tubes and the following panel dials: selectivity, phasing, main tuning, band change, antenna trim and bfo pitch. Next, remove the front panel by removing 11 securing screws, and unhook the two dial lamps over the dial drum. Tilt the panel forward. Now, remove the plastic "kilocycles" dial from the vfo shaft, after first setting the vfo shaft at about the center of the tuning range.

Remove the three screws and lock washers holding the vfo to the chassis panel. Unsolder the two power leads (black/white is filament; red/blue is B-plus to pin 5 of the voltage-regulator tube). Unsolder the coaxial output cable from pin 1 of the 6BE6 mixer tube. The vfo is now free of the receiver.

The easiest way to remove the vfo is to lift it upwards through the top of the receiver, as the under-chassis area is blocked by other components. To do this, remove the bottom tapped panel spacer from the front of the vfo, the spacer otherwise will hit the shaft coupler and prevent the vfo from being tipped forward for removal.

As you remove the vfo, take care to catch and carefully retain the central portion of the flexible shaft coupler, which usually falls out at this point. Note that the upper-right screw holding the vfo, as viewed from the front, is accessible from the front of the receiver through a hole in the tuning gear by turning the "kilocycle" shaft to align the hole over the screw.

Once the vfo is out of the receiver, mount it in the jig. Attach a flexible coupler to the vfo tuning shaft. When you remove the original coupling from the vfo shaft, note its location on the shaft **before** you remove it. This will allow it to be returned to the same position after completion of work. Normally, the coupling plate is flush, or nearly so, with the end of the shaft.

The plastic "kilocycles" dial and an extension shaft permit the vfo to be tuned during tests, and a makeshift pointer is mounted to the aluminum plate so dial calibration may be noted. Place the receiver aside, and cover the main drum dial with a cloth to protect its printed surface.

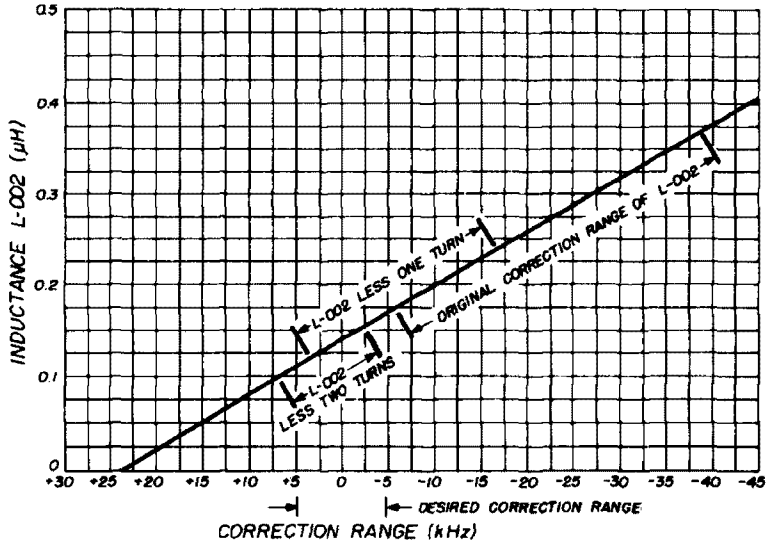


fig. 2. Plot of trimmer coil L-002 inductance versus correction range. Removing one or two turns reduces over-all correction range, but moves it in proper direction to allow correction as shown.

vfo adjustment

Apply filament and plate voltage to the vfo. It should draw about 12 to 15 mA plate current. Make a frequency calibration

chart of the tuning range by checking the 100-kHz points against a well-calibrated receiver tuned to the 2- to 3-MHz range.

Next, remove the vfo shield. The shield will shift the vfo frequency by many kHz when removed. It must be replaced when frequency measurements are made. While the shield is off, however, observe the movement of L-002's slug as you vary it. This will give you a feel as to the amount of expansion or contraction that this adjustment provides. In most cases, you'll find this control won't provide enough expansion for the correction range. A maximum tuning range of about 990 kHz per ten turns of the dial may be found; thus the dial must make about 10-1/8 revolutions for the 1000-kHz range.

Note also that the maximum tuning range is about $12\frac{1}{2}$ turns of the shaft, with the center point about $5\frac{1}{2}$ turns from the 2-MHz end. The most nearly correct tuning range will occur with L-002's slug withdrawn from the coil (minimum inductance).

replacing capacitors

Before you recalibrate the vfo, it's wise to replace certain capacitors. While they don't affect vfo calibration, they're bound to give trouble sooner or later. These are the two 0.01 μF bypass capacitors, C-005, and C-006, located beneath the tube sockets, and the 0.01 μF feedback capacitor, C-008. (The latter is inside the vfo can at the base of the main tuning coil, L-001.) All are "match stick" capacitors, rated at 200 Vdc. They short out frequently, especially when they age. Since the vfo must be removed to make the coil modifications, it's a good idea to replace these capacitors.

Capacitors C-005 and C-006 may be reached by removing the small U-shaped shield beneath the vfo tube sockets. The shield is soldered in place and can be easily removed. Replace C-005 and C-006 with 0.01μF, 600-volt disc ceramics. Resolder the U-shaped shield, making sure the new capacitors' leads don't short to the metal case or cover.

Feedback capacitor C-008 is located in the vfo at the base of the main oscillator coil, L-001. Replace C-008 with a 0.001μF, 300-volt silver mica capacitor. Secure the new capacitor to the adjacent support terminal with a drop of epoxy cement. The vfo is now ready for final calibration.

calibrating the vfo

It's necessary to reduce the over-all inductance of the tuned circuit to achieve the proper tuning range. This is most easily done by altering the inductance of trimmer coil L-002. When new, L-002 has an inductance range of about 0.18 - 0.4 μH (**fig. 2**). In order to bring the adjustment range within practical limits, one turn should be carefully removed from L-002's outer end.

Carefully unsolder L-002's terminal connection, and unwind the wire. Scrape the insulation one turn back from the end. Resolder the wire to the terminal at this point. Replace the vfo's cover, and make several frequency checks at different settings of L-002's slug. If the inductance setting of L-002 is correct, the vfo will tune the 1000-kHz range with ten turns of the shaft and with L-002's slug advanced about six turns into the coil. You should be able to expand and compress the over-all tuning range from 995 to 1005 kHz or more by adjusting L-002. If the vfo was very far out of adjustment initially, it may be necessary to remove an additional one-half turn from L-002 (see **fig. 2**).

The chart of **fig. 2** shows that when L-002's inductance is decreased, its adjustment range is accordingly reduced. With one turn removed, the adjustment range is about 18 kHz; with two turns removed, it shrinks to about 8 kHz. Clearly, no more wire than is absolutely necessary should be removed from L-002, as the point of no return approaches rapidly.

Once L-002 has been adjusted to provide the proper tuning range, the vfo is ready to be returned to the receiver. Firmly bolt the vfo's cover into place. Set the vfo to approximately the center of its tuning range (about 2.5 MHz). Set the "megacycle" dial to an integral megahertz (15 MHz, for example).

table 1. Tuning ranges, operating voltages, and number of tuning dial revolutions to cover tuning ranges of the Collins receiver vfo's.

Type	Use	Tuning Range (MHz)	Filament Voltage	B+	Turns*
70E-1	ARC-2	1.0-1.5	12.6	250	10
70E-2	ARR-15	2.0-3.0	12.6	250	10
70E-3	ARR-15	0.450-0.550	12.6	250	5
70E-7A	75A-1	2.0-3.0	6.3	250	10
70E-8	310B 32V	1.6-2.0	6.3	180-250	16
70E-10	708-A	0.600-0.800	6.3	250	10
70E-11	708-A	1.0-1.5	6.3	250	10
70E-12	75A-2 75A-3	1.955-2.955	6.3	150	10
70E-14	KW-1	1.6-2.5	6.3	210	16
70E-15	51-J	2.0-3.0	6.3	150	10
70E-23	KWS-1	2.75-3.75	6.3	210	10
70E-24	75A-4	1.955-2.955	6.3	150	10
70H-1	R-389	0.469-0.980	6.3	180	51
70H-2	R-390	2.455-3.455	6.3	180	10
70H-12	R-390A	2.455-3.455	6.3	180	10
70K-1	KWM-1	3.455-3.545	6.3	200	1
70K-2	KWM-2	2.5-2.7	6.3	130	2
70K-2	32S-1	2.5-2.7	6.3	130	2
70K-2	75S-1	2.5-2.7	6.3	130	2

*Number of shaft turns to cover vfo tuning range.

The first step is to replace the center portion of the vfo's shaft coupler on the mating coupler section of the receiver. It may be held in place by friction, plus a little heavy grease. Lower the vfo into position with a twisting motion, and when in place with coupling mated, temporarily lock it into position with two retaining screws through the front panel. Replace the bottom spacer (previously removed), and secure the third panel bolt. Rotate the vfo shaft to make sure the coupling doesn't bind. Then return it to an even "megacycle" position on the main dial.

The final job is to align the vfo shaft and the "megacycle" and "kilocycle" dials for correct readout. Tune in a signal of known frequency from a generator or WWV. Various shaft couplings should be slipped until calibration is correct, using the receiver bfo for zero-beat. The process is markedly speeded if the dial setting and various shaft settings are not touched once the original positions are logged before removing the vfo. However, even if the shaft positions are inadvertently lost, no great harm is done. They may be re-established as long as an accurately known frequency is used for final calibration.

A number of Collins receivers, modified as described here, have dial calibrations accurate to within one kHz at any point on the dial. The zero-set adjustment need not be used unless a very accurate frequency readout is desired.

references

1. P. H. Lee, "Further Improvements for the Collins 51-J," CQ, April, 1968, p. 68.
2. W. M. Scherer, "More On Updated Improvements for the 51-J Receiver," CQ, December, 1968, p. 64.
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ham radio

frequency synthesis

A discussion of some digital frequency synthesis techniques for amateur radio operators

L. W. Danz, WA5SKM, Technical Writer, Collins Radio Co., Dallas, Texas

As the amateur bands become more and more crowded and the need for precise frequency control becomes greater and greater, advanced amateurs should abandon the familiar vfo and consider the more sophisticated, more accurate frequency synthesis. The techniques of frequency synthesis have been available for years, but it wasn't until the advent of integrated circuits and computer counting techniques that a practical method evolved.

Forty years ago, a vfo was accurate enough for most communication systems. Then the airwaves became crowded, channels were assigned, and soon (for all except the amateur) the vfo was not accurate enough. Enter the crystal oscillator with its excellent long-term frequency stability. Enter also a cost problem because each desired frequency required at least a separate crystal. This problem was overcome by combining a vfo with a crystal oscillator. When the frequency of a vfo is controlled by a crystal oscillator, the drift of the vfo is constantly corrected so that the long-term drift of the output signal is equal to that of the crystal oscillator itself.

One of the basic elements of frequency synthesis is the phase-lock loop. **Fig. 1** illustrates the basic phase-lock loop in which the output of a voltage-controlled oscillator (vco) constantly is compared with the frequency supplied by the reference source. The reference source usually is a crystal oscillator; the vco may be thought of as a vfo whose output frequency is determined not by a manual dial but by an electrical input.

Because of its long-term instability, the vco tends to drift off frequency. Any change in the output frequency is detected by the phase comparator when compared with the reference signal. When a phase difference exists, the phase detector generates a control voltage which returns the vco to the proper frequency. The long-term stability of the phase-lock loop output is that of the reference source.

voltage-controlled oscillator

The simplified schematic diagram of a typical voltage-controlled oscillator (vco) is shown in **fig. 2.** The components form a common-base Hartley oscillator circuit with L1, C1, C2 and D1 in the tank circuit. R2 and R3 form a voltage divider to establish base bias for Q1; C3 places the base of Q1 at rf ground. The output of Q1 is developed across R1 and L1. Any ac developed across R1 is fed back to the emitter of Q1 through C4 and the tap on L1; this positive feedback causes oscillation. Since the oscillation frequency is determined by the tank circuit, the resonant frequency of the tank circuit can be changed by changing the value of C1 or D1.

D1 is a variable capacitance diode (varactor) which, as long as it is reverse-biased, acts as a capacitance whose value varies in proportion to the dc voltage across the diode. By controlling the dc voltage across the diode, the resonant frequency also can be changed manually by changing the setting of variable capacitor C1.

Capacitor C2 isolates the dc control voltage from the transistor, and the rf choke in series with the dc control voltage, along with capacitor C5, forms a low-pass filter that prevents rf present in the tank circuit from being applied to the dc control lines.

pulse shaping networks

Since this article deals with pulse handling circuitry, this section will review some basic pulse shaping networks shown in ***fig. 3.*** Consider the clipper first. The diode shunted across the output clips the negative excursions of the input pulse (when the diode is forward-biased) but does not affect the positive excursion (when the diode is reverse-biased). Reversing the polarity of the diode would leave the negative-going pulses in the output while removing the positive-going portion of the input waveform.

The clamper shifts the input waveform so that the largest negative excursion (10 for the input) occurs at 0 volts in the output. Thus, the dc component of the output is +10 Vdc, the iargest negative excursion is 0 Vdc, and the largest positive excursion is +20 Vdc. Reversing the polarity of the diode would place the dc component of the output at −10 Vdc, the largest negative excursion at −20 Vdc, and the largest positive excursion at 0 Vdc. Note that the waveform is not distorted.

The differentiating network (or differentiator) is an RC circuit whose time constant is extremely short in comparison with the pulse width (pw) or pulse repetition time (prt) of the input signal. The fast, positive-going leading edge of the input pulse is immediately passed by the capacitor, and all of the output appears across the resistor. As the capacitor charges, more and more voltage appears across the capacitor and less and less across the resistor. When the negative-going trailing edge of the pulse appears at the input, it is immediately transferred to the output by the capacitor, and all of the voltage appears across the resistor.

As the capacitor charges, more and more voltage appears across the capacitor and less and less across the resistor. Thus the input squarewave pulse has been shaped into symmetrical positive and negative spikes coinciding with the leading and trailing edges (respectively) of the input pulse. Either the positive or negative portion of the output waveform could be eliminated with a clipper, leaving one spike for each input pulse.

The ramp-forming integrating net would be a simple low-pass filter if it were not for the diode across the resistor. When the positive-going leading edge of the input pulse occurs, the diode is reverse-biased, and all of the current flows through the resistor. As long as the load on the output is negligible, the current through the resistor is that necessary to charge the

capacitor. Initially, the current is relatively high, and most of the input voltage appears across the resistor. As the capacitor charges, more and more voltage appears across the capacitor, and because less current is required, less and less voltage appears across the resistor.

If the RC time constant is made relatively long in comparison with the prt of the input waveform, the sloping portion of the output waveform will be a relatively straight ramp. When the negative-going trailing edge of the input pulse appears, the forward-biased diode provides an easy discharge path for the capacitor. It also couples the input directly to the output. Since the input is zero and the capacitor is quickly discharged, the output is zero.

phase detector

In the phase-lock loop illustrated in **fig. 1,** the variable output frequency was compared with the signal from the reference source. This is accomplished in the phase detector or phase comparator stage that is illustrated in **fig. 4.** In this circuit Q1 and Q2 are connected as a **flip-flop.** When power is first applied, one transistor conducts more heavily than the other because of natural circuit intolerances. Assume that Q2 conducts more heavily than Q1. The voltage at the collector of Q2 goes less positive, and is coupled through resistor R2 and commutating (speed-up) capacitor C2 to the base of Q1.

The negative-going voltage at the base of Q1 tends to cut off Q1, which results in a positive-going voltage at the collector of Q1. R1 and C1 apply this positive-going voltage to the base of Q2, increasing the forward bias on the base of Q2. Regenerative action occurs, and a stable state is reached with Q2 saturated and Q1 cut off. The output (at the base of Q3) is effectively zero.

If a negative pulse (or the trailing edge of a positive pulse) is applied to capacitor C4 while Q2 is on, diode D2 is forward biased (by the positive voltage on the base of Q2), and the negative-going signal will be coupled to the base of Q2. This pulse momentarily cuts off Q2. The collector of Q2 becomes positive, and this positive-going voltage is coupled by R2 and C2 to the base of Q1, turning on Q1. The resulting low voltage at the collector of Q1 is coupled by R1 and C1 to the base of Q2, holding Q2 off. The output (at the base of Q3) is now positive. Thus, a second stable state has been reached, with Q1 on and Q2 off. Note that Q1 and Q2 cannot both be on or off at the same time.

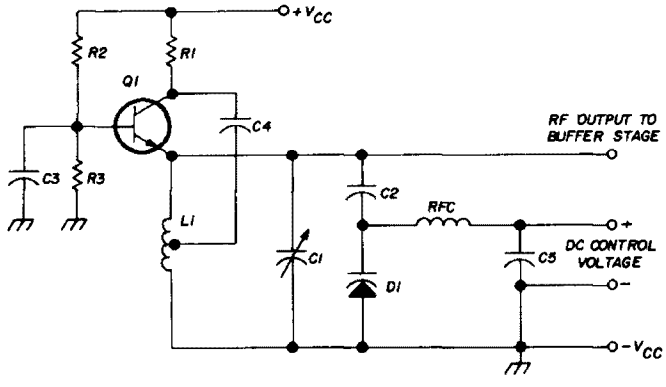


fig. 2. Simplified schematic diagram of a typical voltage-controlled oscillator.

At this point, diode D1 is forward-biased and diode D2 is reverse-biased by the voltages on the bases of Q1 and Q2 respectively. If a negative-going pulse is applied to capacitor C3, Q1 will be momentarily cut off. The collector of Q1 becomes positive, the base of Q2 goes positive, and Q2 turns on. The collector of Q2 goes less positive, the base of Q1 goes less positive, and Q1 is turned off. This returns the flip-flop to its original state: Q2 on, Q1 off. The output (at the base of Q3) is again low.

The timing diagram in **fig. 5** illustrates

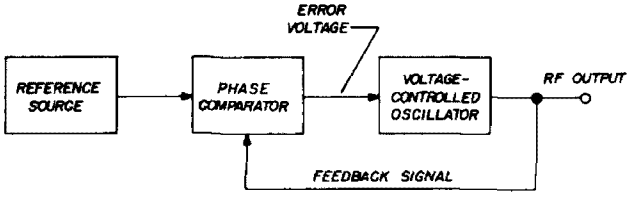


fig. 1. Block diagram of the basic phase-lock loop.

the waveforms present during the application of the flip-flop phase comparator. The reference signal from the reference source is converted into a squarewave by squaring amplifiers (not shown). This input is differentiated by capacitor C4 and the base-emitter resistance of Q2. The positive spike first formed by the differentiator at time T0 has no effect on Q2 since Q2 is already on. The negative spike formed on the trailing edge of the reference pulse at time T2 turns off Q2, and Q1 turns on (as explained earlier). Therefore, at T2 the collector of Q2 goes positive and the collector of Q1 goes to zero and the output is high.

The variable signal from the output of the vco is applied to squaring amplifiers, converted to a squarewave and applied to C3. The positive going spike at T1 has no effect on Q1 because diode D1 is reverse biased by the low voltage on the base of Q1. However, at T2 diode D1 becomes forward-biased by the positive collector voltage of Q2. When the negative spike appears at C3 at T3, it is coupled to the base of Q1, turning off Q1, and Q2 turns on. The collector of Q2 goes low, and the output is low.

The flip-flop is now in its original stable state of Q2 on, Q1 off, and it remains that way until the next negative-going pulse from the reference input at time T6. Then, Q2 turns off, Q1 turns on, and the output is high. The output remains high until the next negative-going pulse from the variable input at T7. This turns Q1 off, Q2 on, and the output is low. The output of the stage is taken across load resistor R5 in the emitter circuit of buffer amplifier Q3. During the time period from T0 to T11, the reference signal leads the variable signal by 90°, and the output pulse has a 25 percent duty cycle.

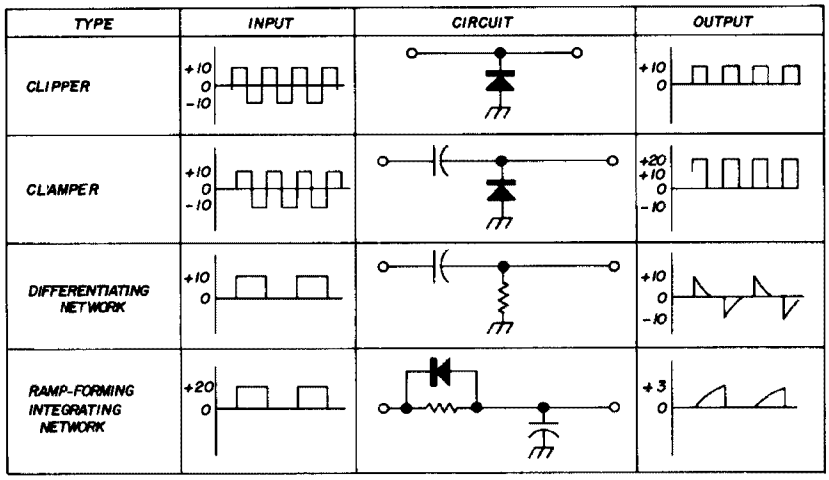


fig. 3. Pulse-shaping networks.

At T11 the phase of the variable signal shifts so that the variable lags the reference by 180 degrees. At T14, the negative-going reference signal turns off Q2 and Q1 turns on. The next negative-going variable signal occurs at T16, turning off Q1, and Q2 turns on. Thus, the output of the circuit is high from T14 and T16 and from T18 to T20, and the output pulse has a 50 percent duty cycle. The pulse width of the output signal is directly proportional to the phase difference between the two input signals and is greatest when the phase difference is 359 degrees. Note that the prt of the output is the same as the prt of the reference signal.

Applying the output from buffer amplifier Q3 to a low-pass filter will provide a dc voltage that is highest when the phase difference is greatest. If this voltage is ap-

plied correctly to the vco, it can keep the vco on the desired frequency.

frequency synthesizer

Most of the elements necessary to construct a digital frequency synthesizer have been covered. One possible configuration is shown in **fig. 6.**

The frequency standard is a crystal oscillator operating at 100 kHz. The 100-kHz output of the frequency standard (Fs) is applied to a fixed frequency divider. It divides input frequency Fs by a fixed devisor (M) to obtain reference frequency Fr. (It will become apparent later that the output frequency of the synthesizer may be shifted in discrete steps equal to the reference frequency. Thus reference frequency Fr must be low to make the steps practical. It is more practical to use a relatively high frequency crystal oscillator followed by a divider than it is to have a properly designed low frequency oscillator to provide Fr directly.)

By changing the divisor (M) it is possible to change the reference frequency, but this change is usually reserved to the design rather than the operational process. For example, if Fs = 100 kHz and M = 10, then Fr = 10 kHz. Or, if Fs = 100 kHz and M = 100, then Fr = 1 kHz. The latter is more practical for application in the amateur bands and is the frequency which will be used in the following discussion.

In a similar manner, the rf output of the synthesizer is divided, but by a variable divisor (N). Assume that the output frequency is 3600 kHz. Then, if N is set (by means of switches on the control panel) to 3600, the output of the variable divider (Fd) is 1 kHz. This 1-kHz signal is compared with reference frequency Fr (also 1 kHz). Any phase difference is detected, and a pulse train is applied to the control voltage (low-pass) filter. The output of the filter is a dc voltage used to control the vco. When divided output frequency Fd equals reference Fr, the loop is "locked."

The beauty of the circuit is that the output frequency (Fo) can be changed merely by varying the divide ratio (N). Assume that N is changed from 3600 to 3590. The frequency of the vco must now change to keep the reference frequency and Fd the same. At the instant of change, Fd equals 1.003 + kHz (3600 kHz ÷ 3590 = 1.003 + kHz).

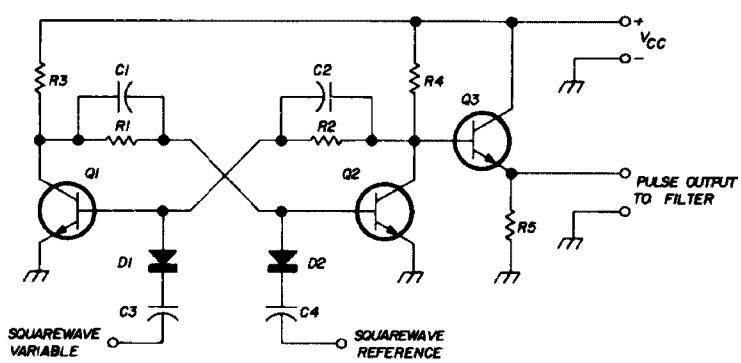


fig. 4. Phase comparator stage electronically compares a varying signal with a stable reference signal.

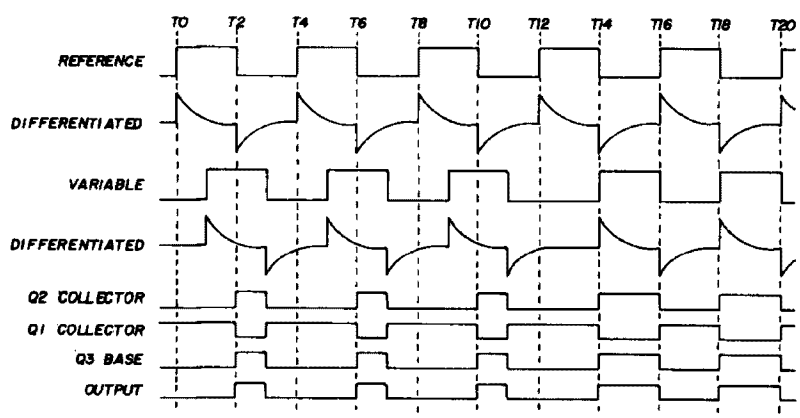


fig. 5. Timing diagram of the phase-comparator stage.

When this value of Fd is compared with Fr the control voltage changes to swing the frequency of the vco (Fo) lower. As the vco frequency approaches 3590 kHz, the control signal continues to change until finally the output frequency is 3590 kHz, Fs is 1 kHz, and the control signal is stabliized.

At this point (at the start of the sequence) the phase detector is not comparing phase but is detecting a frequency **and** phase difference. Without knowing more of the parameters of this system, it's not possible to say that an error of 10 kHz is within the capture range of the phase-lock loop. If the error is outside the capture range,

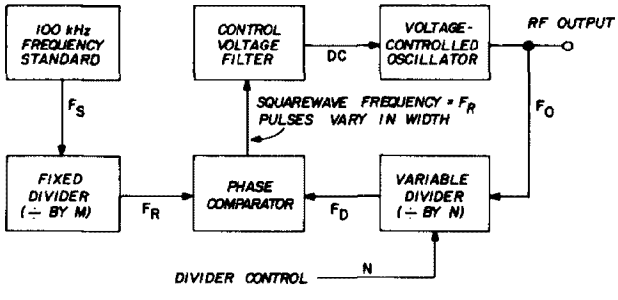


fig. 6. Block diagram of a digital frequency synthesizer.

the loop will be unable to correct the vco output to the new, desired frequency.

A large frequency error may be corrected in two ways:

1. The vco frequency could be changed manually until the output is within the capture range of the system.

2. Another control signal could be generated to rapidly sweep the vco through its entire range. As the vco frequency entered the capture range of the phase-lock loop, the loop would take over frequency control.

The first method is probably the least expensive, especially from an amateur's point of view. It would be a simple matter to monitor the dc control voltage; any control voltage fluctuations which persisted would indicate that the loop was "hunting" and not locked. C1 in the vco could then be varied until the hunting ceased.

frequency comparator

If it's necessary that the synthesizer automatically set itself to a new frequency (as would be the case in a remote control situation), the circuit shown in **fig. 7** could be used. The flip-flop is identical to that used in the phase comparator in **fig. 4.** It is shown here in its logic symbol form, which is easier to work with.

Briefly reviewing its operation, the reference squarewave is applied to the set (S) input, causing the Q output to go high until such time as the variable input signal (applied to the reset [R] input) causes the Q output to go low and the $\overline{Q}$ (read NOT Q) output to go high. (To assist you, the following comparison is made to the flip-flop shown in **fig. 4.** The set input is C4, and the reset input is C3. The Q output is the collector of Q2, and the $\overline{Q}$ output is the collector of Q1.) The frequency comparator in **fig. 7** will provide the same phase difference signal as the phase comparator in **fig. 4.** In both cases the phase difference output is taken through a non-inverting buffer amplifier to minimize loading on the flip-flop.

The frequency comparison circuit makes use of the fact that if a frequency difference exists between the two input frequencies, the higher frequency signal will complete one cycle and at least start another before the lower frequency signal completes one cycle. **Fig. 8** shows the waveforms present in the frequency and phase comparator shown in **fig. 7.**

Note that the frequency of the reference signal (S) is higher than the variable signal. In fact, between times T1 and T7 three cycles of the reference frequency occur while only two cycles of the variable frequency take place. Note also that the reference causes the Q output to be high and the variable causes the Q output to be low. (Remember that only the negative-going trailing edges of the pulses trigger the flip-flop.)

The following is an example of how a frequency difference is detected. The trailing

edge of the reference pulse at T4 causes the Q output to go high. The next trailing edge of the variable input does not occur until between T5 and T6. However, at T5 another reference pulse occurs. The leading edge of this pulse has no effect on the flip-flop (and would have no effect if it was a trailing edge, since the flip-flop is already **set** and the Q output already is high).

The Q output is connected to input E (AND gate A2). The other input (D) is reference signal S. At T5 both D and E are high because Q and S are both high. Therefore, output F (AND gate A2) is high causing gate 3) indicates that a frequency difference exists, but the J pulse is rather short in duration. This pulse could be converted to a ramp-shaped pulse, but the rapid rise-time of the ramp would cause the vco frequency to change too fast to be detected by the phase and frequency comparator. One way to stretch the length of the pulse is to apply it to a one-shot as is done in **fig. 7.**

A schematic diagram of a one-shot is shown in **fig. 9.** Resistor R6 biases Q2 on. R2 and R3 apply a fixed positive voltage to the base of Q1. Since Q2 is on, it is drawing current through R1, and the voltage drop across R1 causes the emitter of Q1 to be

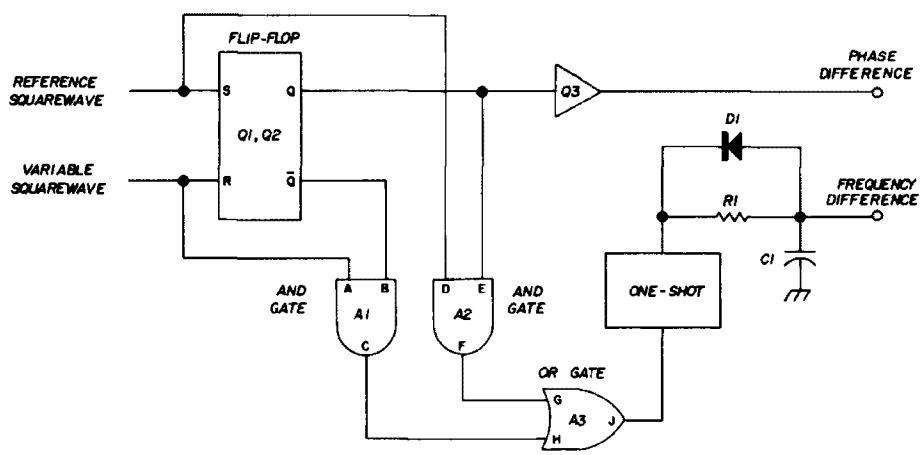


fig. 7. Logic diagram of a frequency and phase comparator.

input G (OR gate A3) to be high. Therefore, output J (OR gate A3) is also high. AND gate A1 is connected in a similar manner to detect simultaneous $\overline{Q}$ and R pulses (which occur when the variable frequency is higher than the reference frequency). In either case, an output from OR gate A3 will occur, indicating that a phase difference of more than 360 degrees exists.

Now that a frequency difference has been detected, it must cause the frequency of the vco to change. One way to do this is to apply a ramp-shaped control voltage pulse to the varactor in the vco. This would sweep the vco through its entire frequency range.

Note in **fig. 8** that the J output (OR more positive than the fixed value of bias at the base of Q1. Thus, Q1 off and Q2 on is the stable state of the circuit.

In this state C2 charges through R4 and the forward-biased base-emitter junction of Q2. A positive trigger at C1 momentarily turns on Q1. The collector of Q1 goes less positive and this negative-going voltage is coupled instantly to the base of Q2 through C2, which turns off Q2. The collector of Q2, and therefore the output, goes high. As time passes C2 discharges through R6, but as long as the base of Q2 is more negative than the emitter, Q2 is reverse-biased and turned off. As long as Q2 is off, Q1 is held on because of the decreased current flow

through R1. When C2 has discharged sufficiently for Q2 to become forward-biased once again, Q2 turns on, and the drop across R1 again reverse biases Q1.

The values of R4, R5 and R1 are critical to the operation of the circuit because the current through Q1 has to be less than the current through Q2. This is so the emitter bias developed by R1 will not turn off Q1 when Q1 is supposed to be on, but will turn off Q1 when Q2 is turned on.

From the preceding circuit description it is apparent that the length of the output pulse is determined by the length of time it takes C2 to discharge through R6. By adjusting the RC time constant of C2-R6 it is possible to adjust the length of the output pulse. In **fig. 8,** the time constant is such that the output pulse is 5 times as long as the reference pulse. In practice it would be many times this length.

The one-shot output pulse in **fig. 7** is applied to the ramp-forming integrating net R1-C1-D1, resulting in the ramp-shaped pulse shown on the last line of **fig. 8.** This pulse is applied directly to a varactor in the vco to sweep its frequency. Once the frequency is within the capture range of the phase-lock loop, the ramp ceases, and the phase comparator takes control of the vco frequency.

fixed divider

One of the elements necessary to the frequency synthesizer shown in **fig. 6** is a fixed divider that divides the frequency of the source (Fs) by 100. Dividing by 100 is accomplished by dividing by 10 and then dividing the result again by 10. Some manufacturers market a divide-by-10 scaler which produces one output pulse for each 10 input pulses. Another way of dividing by 10 is the decade counter in **fig. 10.**

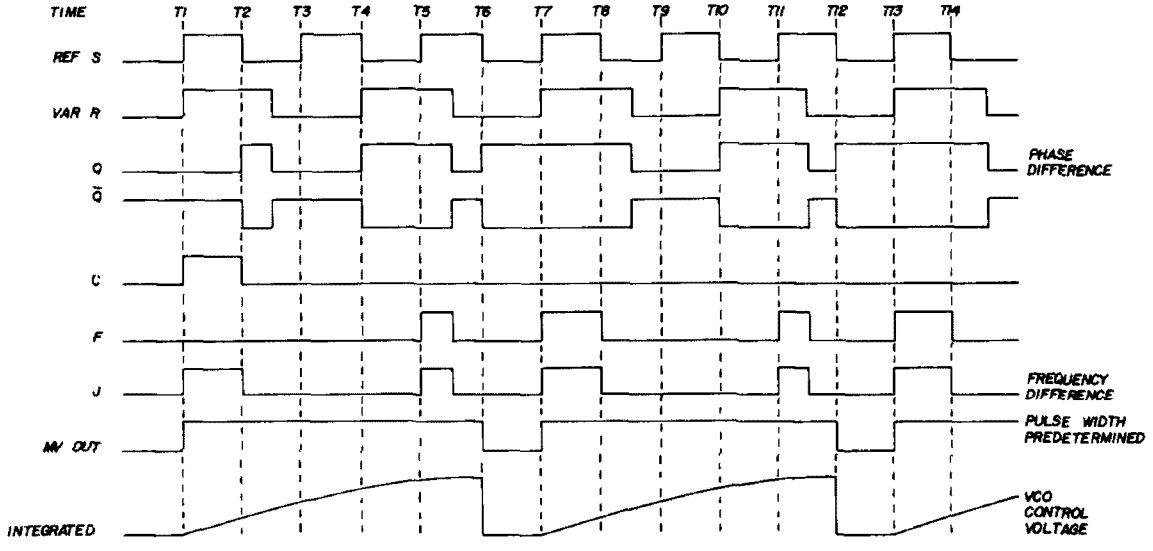


fig. 8. Timing diagram of the frequency and phase-comparator stage in fig. 7.

The flip-flops shown in **fig. 10** are similar to the flip-flops previously discussed (**fig. 4** and **7**) but have an additional input. The toggle (T) input provides symetrical triggering in that a negative pulse causes the Q and $\overline{Q}$ outputs to change state. If Q was high and $\overline{Q}$ was low before the T pulse, then after the T pulse Q will be low and $\overline{Q}$ will be high. The second T pulse again changes the state of the outputs so that Q will be high and $\overline{Q}$ will be low. Applying a string of pulses to the toggle input will yield a pulse train from the Q output with a frequency that is one-half the frequency of the input. A similar out-

put may be taken from the $\overline{Q}$ terminal, but it will be exactly 180° out of phase with the Q signal.

The flip-flops in **fig. 10** may be unsymetrically triggered as before by applying pulses to the set (S) or reset (R) inputs. A negative-going pulse at S will cause Q to be high, while a negative-going pulse at R will cause $\overline{Q}$ to be high. When Q is high, $\overline{Q}$ is low and vice versa. The output conditions cannot be predicted if both S and R are pulsed simultaneously.

In **fig. 10,** the Q output of A (Q_A) is connected to the T input of B (T_B); the Q output of B (Q_B) is connected to the T input of C (T_C). In A, the input pulse is divided by 2; in B, the output of A is divided by 2; and in C, the output of B is divided by 2. This means that the Q output of C (Q_C) is high after four input pulses and goes back to its original state after eight input pulses. Q_C is applied to the set (S) input of flip-flop D.

Note that Q_D is fed back to the R input of B and that Q_A is connected to the R input of D. During the following discussion keep in mind that the flip-flops are triggered only by the trailing edges of the pulses. At the beginning, it is common to assume that all Q outputs are low.

After the first input pulses, Q_A goes high. After pulse 2, Q_A goes low. This negative-going pulse is applied to T_B causing Q_B to go high. (The negative-going Q_A pulse also is applied to R_D, but has no effect because D is already reset). The third and fourth input pulses again cause Q_A to go first high, then low and the negative-going Q_A pulse that occurs causes Q_B to change state from high to low. The negative-going Q_B pulse causes Q_C to change from low to high. Pulses 5 and 6 again cause Q_A to go first high, then low and the negative-going Q_A pulse causes Q_B to change from low to high. This positive-going pulse has no effect on C, and Q_C remains high. Input pulse 7 causes Q_A to go high, but this has no effect on the remaining circuitry.

The action of the flip flops up to this point has been identical to that of a regular binary counter. From this point on the feedback paths play an active part in the operation of the counter. The trailing edge of pulse 8 causes Q_A to go from high to low. This negative-going pulse is applied to T_B and R_D. The negative-going pulse on T_B causes Q_B to go from high to low, and this negative-going pulse causes Q_C to go from high to low. The negative-going pulse from Q_C is applied to S_B. Therefore, there are now two pulses applied to D which seemingly contradict each other (S_D tries to make Q_D high and R_D tries to make Q_D low). Such is not the case, however. The propagation delay of the flip-flop chain removes the contradiction. The R_D pulse is coupled instantly from Q_A but the S_D pulse arrives only after B and C have changed state, so that by the time the S_D pulse arrives, the R_D pulse has lost its punch and the S_D pulse causes Q_D to go high.

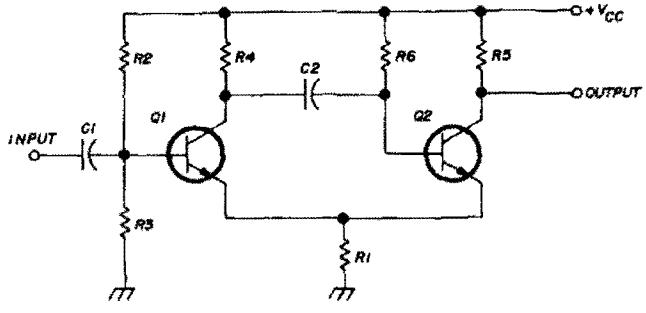


fig. 9. Emitter-coupled one-shot circuit.

Following pulse 9, Q_A changes from low to high, but this positive-going pulse has no effect on the remaining circuitry. After input pulse 10, Q_A goes low, and this negative-going pulse causes Q_D to go low also (because the Q_A pulse is applied to R_D). When Q_D changes from high to low, the negative-going pulse from Q_D is applied to R_B holding Q_B low so that the negative-going pulse at T_B (from Q_A) cannot cause Q_B to change state. Q_C is already low.

After ten input pulses, all flip-flop Q outputs are low, exactly the way they were when the counting started. The same sequence is repeated for the next and all following sets of ten pulses. The output from the decade counter is taken from Q_D. Note

that there is one pulse for every ten input pulses.

Fig. 10 also shows a simple connection to Q_A that provides a divide-by-2 output from the same counter. Note the waveform for Q_A. It contains one output pulse for every two input pulses. The same output could be taken from $\overline{Q}_A$, but it would be 180° out of phase.

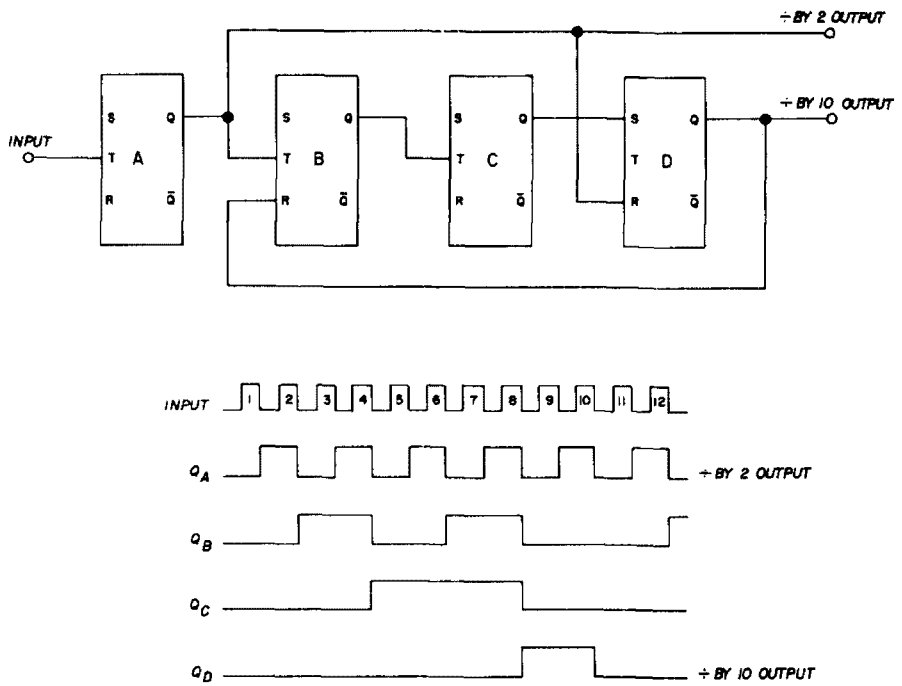


fig. 10. Logic and timing diagram of a decada counter with an auxiliary divide-by-two output.

variable divider

By selectively modifying the feedback paths, it is possible to build counters that will divide by numbers other than 2 and 10, but the design and operation of a variable divider is beyond the scope of this article. It should be noted that the variable divider shown in **fig. 6** should divide Fo by 3500 through 3999 if the output of the synthesizer is to be within the 80-meter amateur band.

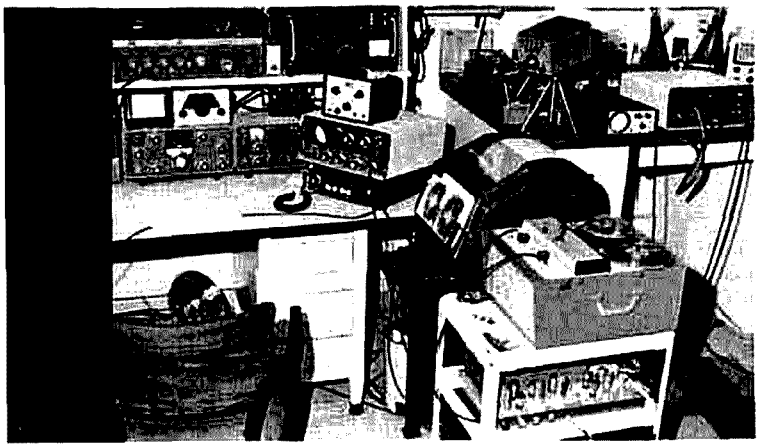
summary

I hope that this article has removed some of the mystique about frequency synthesis. While I have not been able to delve into all of the design considerations, it is hoped that these notes have cleared away enough of the fog surrounding frequency synthesizers to stimulate some "research and development" activity among those amateur radio operators who have cried, "I've built everything there is to build from a spark-gap transmitter to an ssb transceiver." Here is something new to challenge you—I'd be interested to hear what develops.

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ham radio



K7YZZ in Bellevue, Washington is fully equipped for sstv operation.

slow-scan television

The principles of this narrow-band system are discussed, with some ideas on how to get started

Ralph E. Taggart, WB8DQT, 1109K University Village, East Lansing, Michigan 48823

The term "slow-scan television" (sstv) is appearing more and more often in the amateur literature. However, from comments of fellows listening to the slow-scan net, it seems that few realize what it's all about or what can be accomplished with this mode. This article provides answers to some basic questions I've run into, both over the air, and at several meetings where I've exhibited my gear.

what it is

Slow-scan television is a narrow-band image transmission system. A 120-line picture is transmitted every eight seconds. The narrow-band part of the definition is important, because a slow-scan picture requires significantly less bandwidth than a standard phone signal. Sstv is now authorized in Advanced and Extra-Class phone segments of the 80—15 meter bands, the entire phone segments of 10 meters and our vhf and uhf assignments. Every weekend pictures are exchanged between suitably equipped stations throughout the U. S. Pictures have been transmitted between the United States and England, the United States and Antarctica, and between Canada and Sweden.

About the only limitation to the DX capabilities of sstv is the lack of suitably equipped foreign stations. Interest in sstv is rising rapidly in Europe, Africa and Australia, so it appears that a new era of international communications activity is just ahead.

equipment

Present regulations require single sideband for sstv on 80 through 10 meters, while any phone mode may be used in the appropriate parts of the vhf and uhf bands. Standard transmitters and receivers are used for slow-scan work. Collins S-line equipment is used at W8SH, the club station of the Michigan State University where contacts have been made with Canada, Sweden and several states, including Alaska. Other amateurs have been using transceivers with success. The only additional gear includes a picture monitor and camera, with their power supplies and control circuits. The picture monitor is connected to the station receiver audio output, and the camera is connected to the transmitter audio input.

sstv signal composition

Any television system must transmit the following types of information:

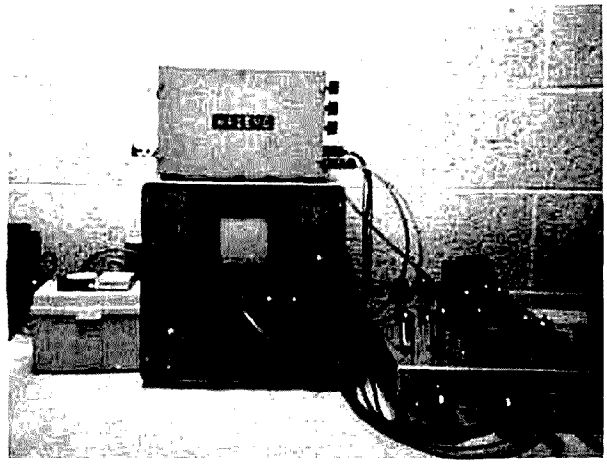
1. Sync signals to ensure that the received picture is in step with the picture being transmitted. These are vertical pulses to start each picture or frame, and horizontal pulses to start each scanning line.

2. Video information, which produces black, white, and gray values during readout of each scanning line.

A single slow-scan frame transmitted on 14 MHz by W4CCG, Manassas, Virginia, and received by W8SH, East Lansing, Michigan. The resolving power of the 120-line system is obviously quite adequate.



The combination of sync and video information comprises the composite television signal. The original series of sstv articles by MacDonald described a system in which the composite signal consisted of an a-m tone or subcarrier.[1] Early experiments, both on the air and in the laboratory, indicated that the subcarrier a-m signal was subject to fading and interference.[2,3] The system was then modified to have a composite signal consisting of an fm audio subcarrier.[3] The equipment was adjusted so that the sub-carrier frequency was 1200 Hz during sync pulses, while the video changed linearly from 1500 Hz (black) to 2300 Hz (white). The vertical sync pulse width was 30 ms and was repeated once every eight seconds. The horizontal pulse width was 5 ms, with a repetition rate of 15 pps. This resulted in a 120-line raster with an aspect ratio of 1:1, which required eight seconds from start to finish (**fig. 1**).



The author's equipment at W8SH.

The subcarrier for a single line scanning a gray scale is shown in **fig. 2.** At A we have the 5 ms pulse of 1200 Hz, which starts the scanning line. The video information ranges from white at B to mid-range gray at C, to black at D, and back to white at E. F represents the 1200-Hz pulse, which starts the next scanning line. Since the 1200-Hz sync pulses are in the infra-black range, the trace is blanked during the retrace interval. With more complex picture material, the sync

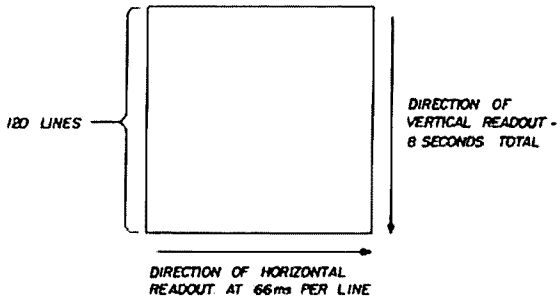


fig. 1. Slow-scan television raster. Aspect ratio is 1:1; readout time is 8 seconds.

pulses would remain the same, but the variations in video information would, of course, be far more complex. The composite slow-scan signal is thus an audio tone that varies in frequency between 1200 Hz and 2300 Hz, a total bandwidth of 1.1 kHz. The signal may be recorded on a standard audio tape recorder, since we are dealing with a mid-range audio tone and not the complex wide-band signal of standard television.

The photograph of a single slow-scan frame shows the resolution of the 120-line system. This signal was recorded off the air using a portable tape recorder with a tape speed of 3-3/4-inches per second. It was then played into the monitor for photography. A cathode-ray tube with a long persistence phosphor (P7) was used in the monitor so that the entire picture could be observed despite the extended readout time.

the monitor

The first step for slow-scan operation is to build a monitor for picture display. The most commonly used circuit was described by MacDonald.[4] My monitor, shown in the photo, was built from this article. Cohen[5] described an economical approach to the construction of this monitor using a surplus radar tracking set. This has been used successfully by many amateurs.

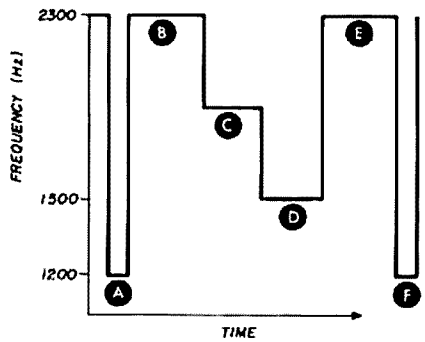
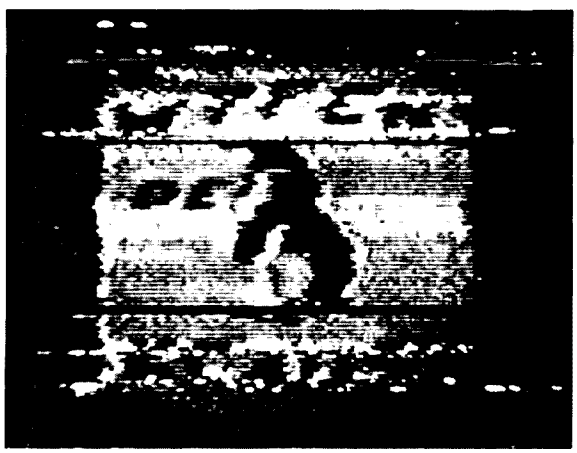


fig. 2. The fm subcarrier frequencies as functions of time. Extremes of tone values, white to black, occupy a bandwidth of only 1:1 kHz.

There's nothing particularly critical about monitor construction, since only audio frequency signals are processed in the unit. Well-filtered power supplies and shielded leads in input circuits will prevent hum pick-up, which is probably the biggest single factor to be avoided. Most of the parts, including the cathode-ray tube, are available from surplus.



Single frame of a transmission from KC4USV, Antarctica, as received by W7FEN. The echo on the right side of the penquin is due to multipath propagation.



Single frame from a 20-meter transmission from W7FEN. W7FEN was using independent sideband at the time with the picture on the upper sideband and the voice on the lower.

Once a monitor has been completed, it is best aligned using a pre-recorded slow-scan tape. I'll be happy to provide such a tape to anyone actually constructing a monitor. Most operators will also make simultaneous recordings of incoming signals to provide a permanent record. It can be played back as often as you wish. The network of **fig. 3** will allow visual and aural monitoring as well as recording.

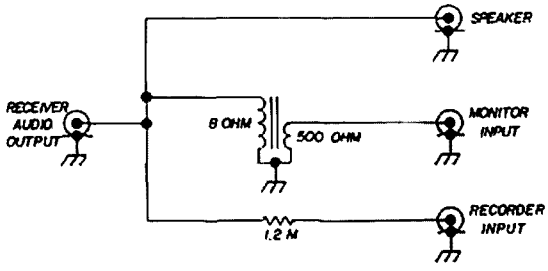


fig. 3. Circuit for visual and audio monitoring of sstv signals. Receiver i-f pass-band should be at least 2.5 kHz to avoid distortion caused by interference.

single sideband operation

When receiving pictures transmitted on single sideband, carrier insertion must be as precise as possible if the subcarrier frequencies are to be reproduced accurately. Small errors in carrier insertion will result in slight shifts in the gray scale, while large errors will usually mean no pictures at all due to the inability of the monitor to trigger properly on the sync pulses. Proper carrier insertion is easy if the monitor is in use, but may be a problem if you are recording "blind" for later playback. In this case, it is best to wait for a voice transmission from the station and tune for the most natural-sounding voice quality. Although this may sound like an indefinite procedure, it is surprisingly easy and will almost always be close enough for good picture rendition.

A signal of S4 or S5 on the hf bands will produce a readable picture if a reasonably clear channel is available. Signal levels higher than S9 will give "tape quality" pictures in the absence of strong interference. Since interference is the limiting factor in hf reception, the receiver should have a narrow i-f response; not more than 3 kHz, and 2.5 or 2.1 kHz is even better. If the receiver i-f response is too broad, an external audio bandpass filter with cutoffs at 1 and 2.5 kHz would be a worthwhile addition.

vhf and uhf operation

Operation in the vhf and uhf bands is also feasible, but here the limiting factor is usually the signal-to-noise ratio rather than interference. The use of a-m or narrow-band fm on vhf eliminates the carrier insertion problems associated with ssb, but just as in normal phone operation, the use of ssb will result in a sizable increase in reliable range. I've been successful in transmitting sstv pictures about 100 miles on two meters using just 75 watts of a-m, so this type of operation should not be overlooked.



On-the-air shot of W9NTP. Hare Don is adjusting the slow-scan monitor. The shuttered camera above the operating position is normally operated in real time.

Once the receiving gear is set up and working, the urge develops to send pictures. There are two general types of picture generators now in use—the flying-spot scanner (fss) and the vidicon camera. The fss is suitable for transmitting from photographs and other printed material. An excellent unit of this type has been described by Hutton.[6] A number of operators are presently using units of this type. A more versatile approach is a live vidicon camera, described by MacDonald in a three-part article in QST.[7] This is uses the first of the two approaches, but both are highly satisfactory.

Although some live transmissions are made, most operators record their pictures for later playback over the air. This technique permits a maximum variety of material to be transmitted in as little on-the-air time as possible, since the tape may be edited as it is put together. Some provision must be made to feed the source (camera or recorder) into the transmitter audio input. The circuit of **fig. 4** is useful for this purpose.

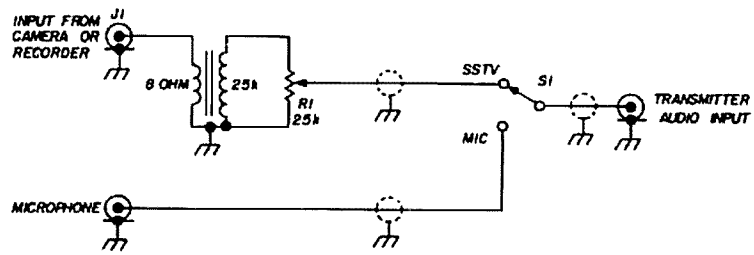
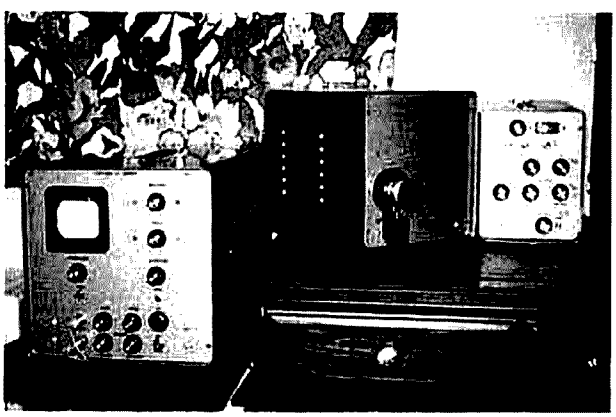


fig. 4. Schematic for feeding video and audio signals to the transmitter. With recorder or camera level preset, video and audio input can be switched without changing transmitter audio gain control.

probably the most commonly constructed camera to date and gives very good results.

As originally described, the camera uses a special slow-scan vidicon, which is no longer available on the **gratis** basis described in the original articles. To get around this problem, many amateurs have experimented with standard vidicons. The procedure described by Taggart[8] requires no modifications of the original circuit, while that of Hutton[9] involves changes in the vidicon and video amplifier circuits. My camera (in the photo)



Slow-scan tv equipment used by W4UMF includes the hybrid monitor to the left and the transistorized vidicon camera in the center.

The transmitter audio gain control is set with the microphone at your normal speaking level. With the recorder or camera feeding J1 at some preset level, S1 is placed in the slow-scan position, and R1 is adjusted for the same level as when the microphone was used. If no changes are made in the recorder or camera level, you can then switch back and forth between voice and picture with no changes in the transmitter gain control. During the first few attempts at transmitting sstv, it is best to keep an eye on the final amplifier plate current. The constant amplitude of the fm subcarrier runs the final at a pretty stiff duty cycle, and a blower or slightly reduced power level may be required if the final tubes are gassy or the transmission line standing wave ratio is high.

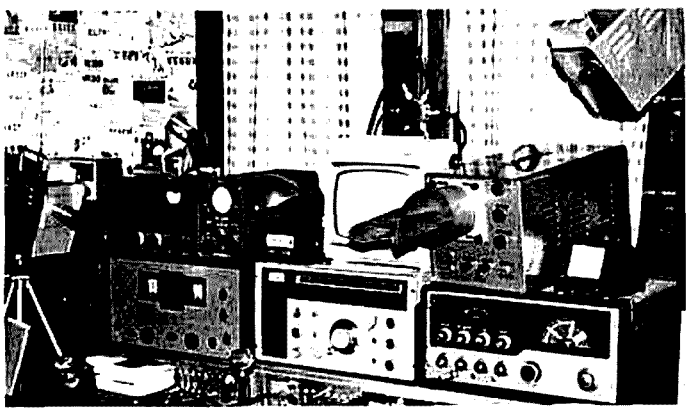
sampling techniques*

Many amateurs engaged in sstv formerly experimented with wideband tv on the 432 MHz band, so they already had vidicon cam-

*This material on sampling techniques was originally presented to the FCC in a sstv progress report, "Slow-Scan Television Activity in the Alexandria, Virginia, Area," 27 September, 1968, by T. J. Cohen, W4UMF, D. Harmon, W4CCG, and D. Veazey, W4ABY.

eras which produced standard non-interlaced pictures suitable for conventional television receivers. If they could use these cameras in a slow-scan system it would provide an economical and versatile approach to sstv. Indeed, the following idea was quick to take form: if some way could be found to **sample** different portions of successive fast-scan pictures, the conventional video sources already available could be used as part of a slow-scan camera.

The first amateur slow-scan sampling system was designed and built by Don Miller, W9NTP, in 1966-1967. This system, which used a vidicon camera as a video source, sampled 120 successive fast-scan pictures to produce one slow-scan frame. The sampled pictures, which were first aired on the Navy MARS sstv net in 1967, were judged to be as good as those produced by sstv cameras using special slow-scan vidicons[7]. Miller's circuit design was made available to the amateur community in late 1967; we built and operated a system based on this design in early 1968. Complete circuit details for Miller's sampling camera have been published by CQ.[9,10]



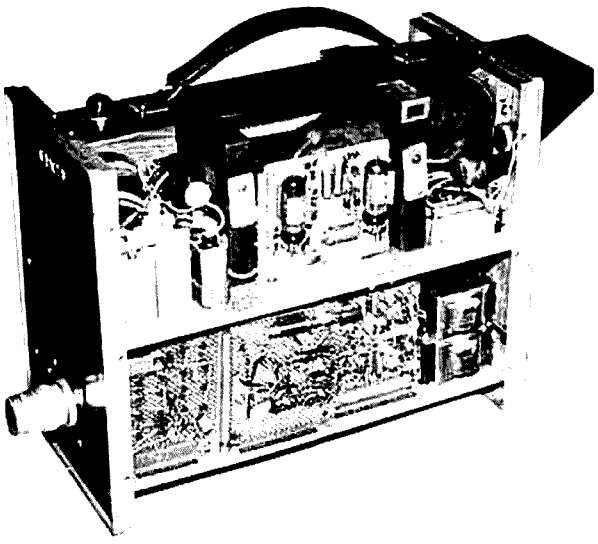
Don Miller, W9NTP, is wall equipped for sstv work. Equipment includes the sampling slow-scan camera to the left, tho shuttered camera, upper right, and both slow- and fast-scan monitors.

Recently, Watson and Horne[11] described a slow-scan sampling camera designed by the Research and Developments Labs, Northern Electric Co., Ltd. (Canada). Though little is known of this system, it works on the same basic principle as Miller's camera: different portions of successive fast-scan pictures are sampled to produce a slow-scan frame.

Sampling-type sstv cameras are now used by at least three amateurs (WB6ZYE, W9NTP and VE3EGO), and construction is contemplated by more. With the growing interest in sampling systems, the need arises for a brief, but lucid, review of sampling techniques.

The block diagram for a sampling-type slow-scan camera is shown in **fig. 5.** A conventional fast-scan camera (with a fast-scan vidicon) is used as a video source. However, as seen in **fig. 5** a line-locked 15 Hz sawtooth waveform is generated in the system and imposed on the vertical deflection coils of the vidicon yoke (the 15 Hz oscillator also generates the slow-scan line sync and blanking pulses). The fast-scan video output now has a frame frequency of 15 Hz; the line frequency remains unchanged at 15,750 Hz.

The modified fast-scan video is fed to the sampler circuit which acts as an electronic gate, "opening" and consequently "sampling" the video at times dictated by the output of the Schmidt trigger sliding-pulse generator. This sliding-pulse genera-



Sampling camera built by W9NTP. The upper level houses a P7 monitor for both fast-scan and sampled slow-scan tv; the lower level is a fast-scan camera with a sampling circuit to provide slow-scan output.

tor is driven by an 8-second ramp waveform and the 15,750 Hz pulses from the fast-scan camera. The resultant output of the Schmidt trigger is a sliding pulse which moves with an 8-second period and has a 15,750 Hz rate.

Once sampled, the video appears as a series of spikes, with the amplitude of each spike equal to the amplitude of the fast-scan video signal at the instant the gate was opened. These pulses are very short in duration, however, and thus it is necessary

tures are shown diagramatically in **fig. 6.** The sampling action starts in the upper left corner of the fast-scan picture. Every 1/15,750th of a second, roughly, a video sample is taken, the effect being to sweep the image downward and to the right (remember, the sampling pulse is sliding in time). Since the vertical frame rate is 15 Hz, 1050 samples are taken before the sampling pulse returns to the top of the picture. Each return is slightly to the right of the previous scan, and in this way, the

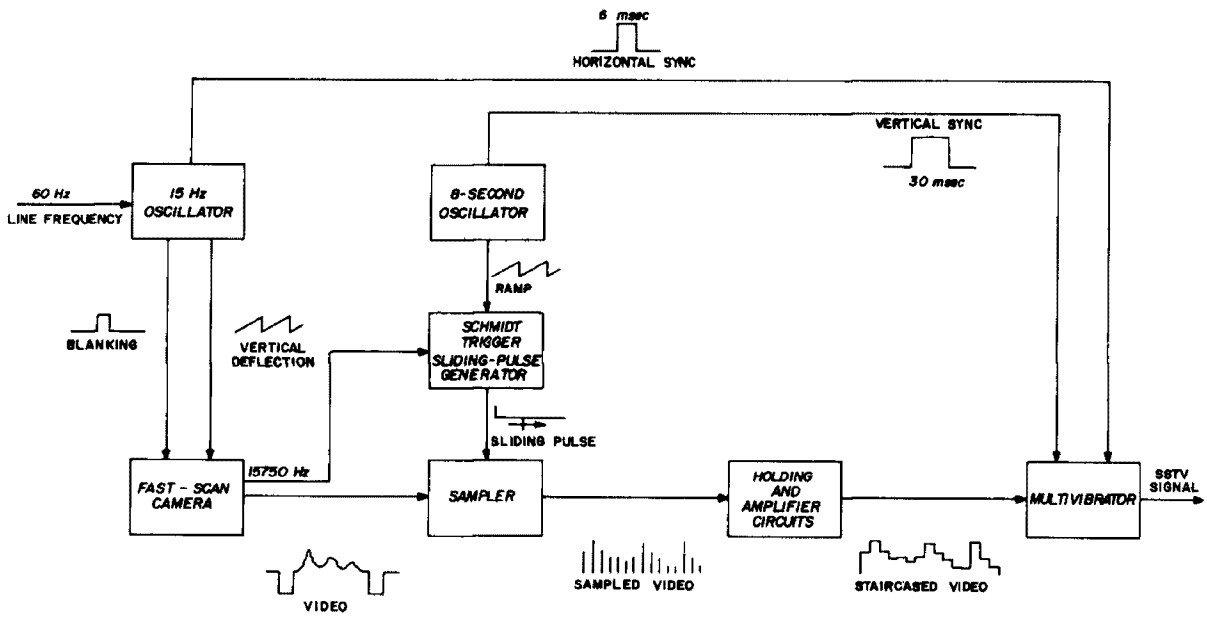


fig. 5. Block diagram of a sampling-typo slow-scan television camera.

to "hold" a given sample at its peak amplitude until the next spike appears. The holding circuit performs this function, and its output, the "staircased" video, is amplified and fed to the multivibrator. Slow-scan horizontal and vertical sync pulses are also fed to the multivibrator, and together with the video input, a composite slow-scan signal is generated with the following characteristics:

1200 Hz	Sync
1500 Hz	Black
2300 Hz	White

Gray levels generate signals between 1500 and 2300 Hz.

Let's examine the sampling method in detail. Fast-scan and sampled slow-scan pic-

entire fast-scan picture is eventually sampled. With a vertical frame rate of 15 Hz and a sampling pulse which slides with a period of 8 seconds, the sampled slow-scan picture consists of 120 lines and is made up of successive portions of 120 fast-scan frames.

When the slow-scan picture is displayed, the 15 Hz frequency is now the line frequency and the vertical frame period is 8 seconds. This results in a sampled slow-scan picture which is both rotated and inverted. However, this presents no problem. For fast-scan cameras with identical vertical and horizontal yoke windings, it is only necessary to interchange (with a switch) the horizontal and vertical deflection leads, and

reverse the "new" set of vertical deflection leads.

With this sampling technique a conventional vidicon camera can be used to produce slow-scan television pictures. Though the subject to be televised must be held fixed for the entire 8-second sampling period the slow-scan pictures produced by a sampling system are of comparable quality to the excellent images produced by cameras using slow-scan vidicons (e.g. the WL-7290). The fact that the sampler camera uses a fast-scan vidicon and can be tuned up in "real time" are but two of the reasons why sampling systems are becoming so popular among sstv enthusiasts.

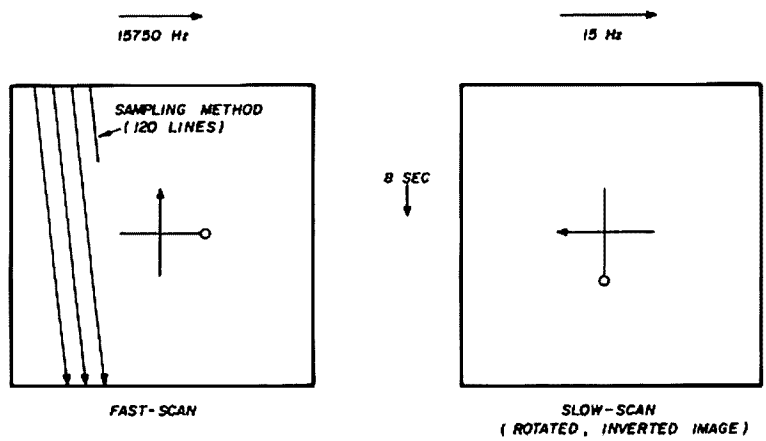


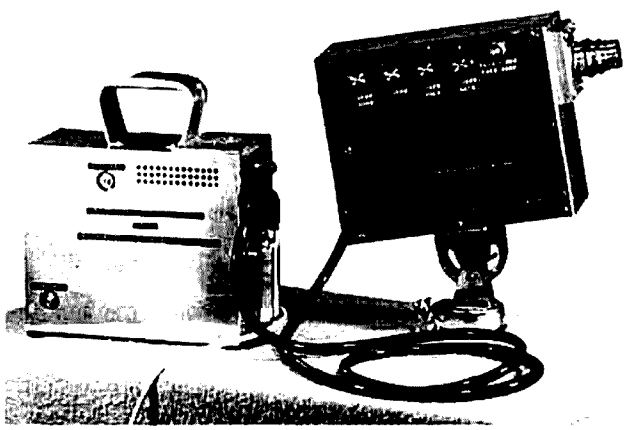
fig. 6. Fast- and slow-scan images.

slow-scan color tv*

Recent color transmissions from the Apollo spacecraft have once again focused attention on the field-sequentional system for transmitting color television pictures. This system, developed by Dr. Peter Goldmark of CBS Labs, and used for the first color tv broadcast in 1940, was rejected by the FCC for the United States standard. However, simplicity of the system (it requires only one-third the components required by a conventional color tv camera) results in reduced camera size and weight and makes it ideal for many medical, industrial and space applications.

In the field-sequentional system of picture transmission, a color wheel containing red, green and blue filters is spun in front of a black-and-white television camera. The resultant black and white pictures, in sequence, contain the color information as seen through the filters. These black and white pictures are called the separation pictures. If the separation pictures are displayed on a crt and viewed through a camera-synced color wheel, the persistence of vision of the human eye will cause the original picture to be reproduced.

On February 7, 1969, Theodore Cohen and Wade Tarr succeeded in reproducing the first pseudo-field-sequentional color picture transmitted by slow-scan television; color synthesis was achieved with subtractive[1] synthesis techniques. Working independently, WB8DQT reproduced the first additive-synthesis color sstv picture on February 13. Further experimentation by Cohen and Tarr led to a method of sstv color transmission which used quick-process-



Solid-state slow-scan sampling camera built by Syd Horne, VE3EGO, and designed by Doug Watson.

*This section on slow-scan color tv was prepared by Ralph Taggart, WB8DQT, Theodore J. Cohen, W4UMF and Wade L. Tarr.

ing Polaroid color film for both the color analysis* and color synthesis processes.

field-sequential color systems

The field-sequential method of color transmission is shown diagramatically in **fig. 7.** The subject to be televised is a black card containing three circles—one red, one green, and one blue (the additive primary colors). These circles are viewed by a black-and-white television camera, the light first having passed through a spinning color wheel. The wheel contains wedges of red, green and blue filters. If the color wheel is synced to the black-and-white camera's frame rate such that successive frames are viewed through successive filters, the first frame in a three-frame set will contain that color information which passes, say, through the red filter. (Note that it is not necessary to synchronize the color wheel to the frame rate—it is done here only to simplify the explanation of the sequence of events. In usual practice the rotation of the color wheel is greater than the camera's frame rate.)

Since a filter passes light of its own color, only the light from the red circle will image on the camera's pick-up tube. This will be televised as a white circle while the other circles and the black background will be televised as black areas. Following in sequence will be a frame containing a white circle in the green circle's position, and a frame containing a white circle in the blue circle's position. The three-frame sequence constitutes a **color analysis** of the color subject. Since color analysis information is transmitted in sequential frames, this system of color transmission is called the field-sequential system. (The present U.S. color-tv standard uses a dot-sequential system).

Upon receipt of the black-and-white color-separation pictures, the images are displayed on a conventional cathode-ray tube. However, the pictures must be viewed by the human observer through another spinning color wheel. If the viewing color wheel is synced to the color wheel used in the color-analysis phase—and they must be synced to reproduce the color image—an observer's eye is presented with the following sequence of color pictures: frame 1, red circle on black background; frame 2, green circle on black background; frame 3, blue circle on black background; frame 4, same as frame 1, etc.

Because of the persistence of vision of the human eye, each frame is stored in the eye for a fraction of a second, and if the three-frame sequence is viewed in rapid succession,

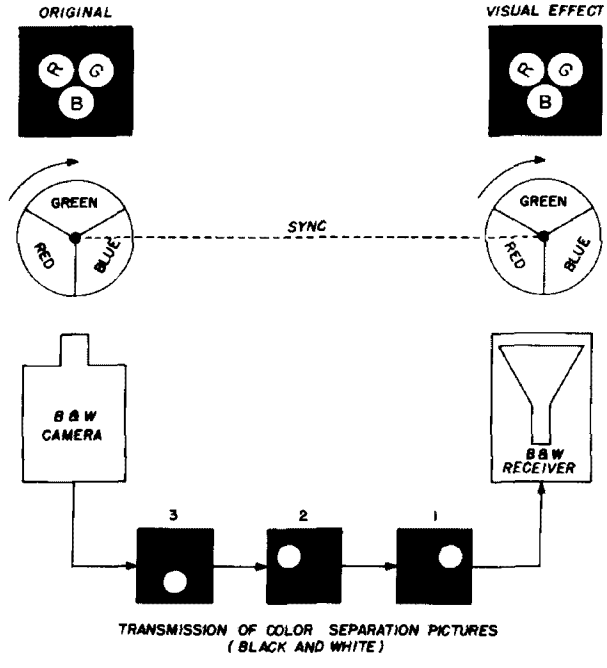
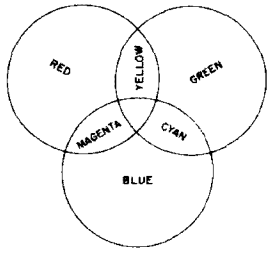


fig. 7. Field-sequential method of color transmission.

*In 1861 James Maxwell demonstrated the principles of three-color photography. He analyzed a landscape through red, green and blue filters, and made three black-and-white lantern slides. Then he projected the slides through red, green and blue filters and produced a color picture of the landscape by superimposing the three pictures. Producing a color image in this way is called **additive color synthesis.**



When white-light sources are projected through red, green and blue filters, and allowed to overlap as shown at the right, white is produced where all three colors are superimposed. There are also three other colors produced: cyan (white light minus red), magenta (white light minus green) and yellow (white light minus blue). This suggests that color synthesis may be performed by **subtractive** mixing using cyan, magenta and yellow. Subtractive color synthesis is the basis of our present color photography and color printing processes.

the observer sees the original color picture. That is, the eye acts as a momentary storage device and integrates the three primary color images into a composite color picture.

At this point, you're probably wondering how a color picture is actually transmitted by slow-scan television. After all, slow-scan tv reads out one frame every 8 seconds, hardly short enough for the eye to integrate three frames into one color picture. The answer to this apparent paradox is that while three sstv separation frames are transmitted in sequence, **color film**—not the eye—is used to synthesize (integrate) the pictures.

Although the first slow-scan color picture was reproduced using subtractive synthesis, later experiments indicated that additive synthesis is more applicable to amateur work. Two basic additive synthesis methods have been investigated to date. One method requires a darkroom and is rather complex; the second method uses direct-write techniques, and can, with the use of high-speed color film, produce color pictures directly off the sstv monitor.

This method of transmitting and receiving color pictures can be broken down into three distinct phases: analysis, transmission and synthesis. In the analysis phase the original color pictuer is viewed sequentially through red, green and blue filters; the three black-and-white separation pictures obtained contain all of the information required to reproduce the original scene. In the transmission phase the separation pictures are transmitted in a clearly defined sequential fashion. Finally, in the synthesis phase, the original scene is reproduced by sequentially viewing the set of separation pictures through primary-color filters.

sstv color analysis

The purpose of the color analysis phase is to yield a set of black-and-white separation pictures. If the television imaging system has a flat spectral response in the visible region, separation can be achieved directly by taking three successive television pictures of the color object through each of the primary-color filters. Although the imaging systems used by amteurs do not have absolutely flat response, flying-spot-scanners using 931A photomultiplier tubes and sstv cameras using conventional fast-scan vidicons exhibit satisfactory response characteristics.



These black-and-white color-separation prints show the red, green and blue (from left to right) components of the original scene.

For indirect color analysis a camera loaded with black-and-white panchromatic film is used. Both 35 mm and Polaroid cameras have been used for this purpose. Again, three separation pictures must be taken—one each through a red, a green and a blue filter. After the three black-and-white separation pictures have been printed, they may be televised using any sstv imaging system. The pictures are then recorded on magnetic tape for subsequent transmission.

sstv color transmission

The end result of the analysis phase is a recorded tape with ten or more frames each

of the red, green, and blue separation pictures (again, these are black and white pictures). Since the sstv sync and video information is all within the audio bandpass, the pictures may easily be transmitted with an ssb transmitter or over telephone lines. Alternatively, the video tape can simply be mailed to the station at which the synthesis is to be performed. As long as the operator at the receiving end knows which frames represent the red, green and blue separations, it is possible to synthesize the color picture. Because of the many monitor and camera adjustments required during synthesis, it is not practical to attempt it in real time (e.g., as the pictures play out from a ssb transmission).

sstv color synthesis

In direct-write synthesis, a 33 mm camera loaded with high-speed color film such as Ektachrome (ASA 160) is mounted rigidly in front of the sstv monitor; a suitable close-up lens is used on the camera. The room should be dark and if a yellow or orange filter is used on the crt face for normal viewing, it should be removed. A single "red" frame is exposed through the red filter, a "green" frame is exposed through the green filter and a "blue" frame is exposed through the blue filter (without advancing the film). The final composite picture will be an excellent rendition of the original color object if the procedure was done with reasonable care.

Polaroid Polacolor has been used for direct-write color synthesis, but because of its slower speed, a large number of exposures must be made through each of the primary-color filters. Because of sync jitter, however, an unavoidable loss of detail results from superimposing a large number of sstv pictures. The advantage of Polacolor, of course, is the ability to rapidly see the results.

summary

This, in very short form, is the story on slow-scan tv, one of the more challenging areas in amateur radio today. Even if you never decide to try it, you will know what's going on when you run into it on the amateur bands. Interested amateurs are invited to check into the slow-scan net (14230 kHz, 1900 gmt on Saturdays), or you can just listen in to hear what it sounds like. Don't hesitate to make tape recordings, for all slow-scan stations are interested in how their pictures look and will provide comments and sometimes photos of the picture readouts if you send them the tape. Three and three-quarter ips is the speed most commonly in use for tapes, and most stations can handle this speed. Finally, don't hesitate to write if you have questions or problems. Slow scanners are a friendly bunch and welcome all interested parties. Join the fun, and see what the fellow at the other end **looks** like for a change!

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Readers who want more information about slow-scan television will be interested in *ATA International*, the international English language edition of the Amateur Television Association; $3.00 per year from Willy P. Everaert, ON4WM, Park Ten Hove, 97 Melle, Belgium.

ham radio

notes on directional swr indicators

Walter Schreuer, K1YZW, Riverbank Road, Ipswich, Massachusetts 01938

Much has been written on the subject of reflectometers (sometimes called directional watt or power meters, relative power meters, etc.) yet considerable misunderstanding of these basically simple devices still seems to exist; fallacies are often overheard on the air and even seen in print. I hope the following notes will help to clear up the more common misconceptions.

some basic facts

All directional instruments, from the least expensive relative-reading units to the most costly absolute meters, have one fact in common: both forward and reverse powers are measured, and the combination is used to determine the standing-wave ratio. Each instrument is set up for a particular **resistance,** which is often erroneously called impedance. Any swr indicated by (or calculated from) the meter readings is a measure of the impedance present at the output terminal of the instrument relative to its design resistance. Note that the impedance of the power source on the input side of the meter has absolutely no bearing on the measurement.

If you put a 100- or a 25-ohm resistor across the output terminals of your 50-ohm swr meter, it will indicate 2:1. However, there are an infinite number of resistance-capacitance and resistance-inductance combinations which **also** produce an indication of 2:1. The standing-wave ratio is not a measure of the nature or magnitude of the impedance, but rather of the quality of the match to its design value—or the severity of the mismatch.

transmission-line impedance

The most common use of a reflectometer is to determine the swr of an rf transmission line. However, you can get meaningful results only when the characteristic impedance of the cable is the **same** design resistance of the reflectometer. When the characteristic impedance of the line is the same as the design resistance of the reflectometer, the swr indication will be the same wherever the reflectometer is placed in the transmission line as long as line loss is negligible. This is usually the case on the high-frequency bands with modern coax. When coaxial cable loss is not small, the swr indication will increase steadily as you move the instrument closer to the antenna.

If you must use your 50-ohm instrument in, say, a 90-ohm system, and you get a reading between 1.6 and 2.0, add a quarter wavelength of transmission line. If the reading does not change much, the actual swr is near unity.

For widely different indications the exact cable length has to be known or a large number of readings taken with different line lengths temporarily inserted to permit calculations of the true swr. When the instrument is moved toward the antenna in a line whose impedance differs from the instrument design value, the indications will change up and down although the **true** swr is the same at all points. This can be veri-

fied by laborious computations or with an instrument designed for the particular cable impedance.

effect of power level

With the simpler types of reflectometers it is quite common to observe a lower swr with low power than with high power. Also, this effect is usually more pronounced at the lower frequencies where the meter is least sensitive. This is because the solid-state diodes which convert sampled rf voltages to dc to operate the meter exhibit a threshold effect: they require a minimum signal before rectification can take place.

This fixed threshold level is insignificant for the rf voltages used for measuring forward power, but it can result in erroneous reverse readings at low power levels. In fact, the results will be most nearly correct for high power levels. Unfortunately, this goes against the usual advice of using minimum power when tuning up antennas.

In the more costly instruments, sensitivity does not change with frequency, and the meter scale can be calibrated directly in watts. With these instruments it is possible to partially compensate for diode threshold by increasing the sensitivity of the reverse power indications over the corresponding forward power ones. Each compensation is correct for only one exact reverse power level (or swr value with one particular forward power), but it is a fair compromise for other levels and a lot better than no compensation at all.

As the designer of Comdel's DW1550 directional wattmeter I found no need for compensation above 50 watts forward power. I chose an swr of 2:1 with a forward power of 15 watts (full scale of lowest range) as the point of exact compensation. For swr measurements with forward powers between 15 and 150 watts (full scale of the next higher range), the effect of the compensation is reduced automatically by the variable resistor which is used to set the meter needle to full scale.

There is an amusing side effect to this method of compensation. When a drastic mismatch occurs, such as an open feeder—hopefully at low power—the reflected power indication with its increased sensitivity will be larger than the forward reading. In other words, the instrument is overcompensated since the reverse power is much higher than the value of exact compensation. This peculiarity is of little practical consequence since in reality reverse power can obviously never exceed the forward power.

using the instrument

Most reflectometers absorb insignificant amounts of power and may be permanently left in the line unless the swr is exceptionally high. All instruments dissipate a small amount of power and may be subjected to damaging high voltages when using a full kilowatt input with a high standing-wave ratio. Manufacturers' data sheets seem to ignore this point; if you are in the habit of using high power with an swr greater than 3:1 you should ask the maker of your instrument for advice.*

When a directional instrument is permanently installed in the feedline, it makes a very useful monitor as well as a transmitter tuning aid. With an inexpensive relative-power indicator the sensitivity is different for each band, but a pointed knob on the sensitivity control with a mark for each band on the panel will overcome this effect. Absolute-reading instruments are more convenient, but they are at least four times as costly.

I invariably tune up my linear amplifier by forward power indication; an occasional check of plate current always shows it to be at or very near its proper value. On a few occasions in the wintertime, my swr meter has prevented possible damage to the linear in the early morning hours when it gave abnormal indications due to antenna icing.

When you are adjusting the matching of a new antenna, and you are sure of the quality and impedance of the feedline, the reflecto-

*The Comdel DW1550 will stand a 4:1 swr at 1.5 kW forward power continuously and higher values intermittently. The Collins 302 C3 wattmeter (and wattmeter in the 312B4) will stand 2 kW continuously at any swr. The Drake W4 and WV4 are limited by the full-scale reading of the meter movement. That is, the sum of the forward and reflected powers should not exceed 2000 watts; for 1500 watts input this occurs at an swr of 3:1.

meter may be located anywhere in the line. If you are not certain of your cable or are using a temporary one with a different impedance, then put the instrument right at the antenna. If this is impossible use an exact multiple of an electrical half wavelength between the antenna and meter.

When checking the load impedance seen by the transmitter, the reflectometer belongs right at the transmitter's output socket. Usually the swr will be the same at any other point in the outgoing rf cable, but if you have an external low-pass filter for tvi elimination as I do you may be in for some surprises on your highest frequency. This is because the more simple filter types show wide impedance variations near the cutoff frequency.

When building my 10- and 15-meter three-element quad, I observed the following readings at 29 MHz: 1.5:1 at antenna, 1.4:1 at the bottom of 100 foot RG-8/U cable, and 2.1 at the transmitter ahead of the low-pass filter. At 21.3 MHz the swr was essentially 1.2:1 at the same three points. As long as transmitter tuning and loading adjustments can deal with the increased swr caused by the low-pass filter, the impedance change is of no practical importance.

Antenna-tuning units are often used for resonating wire radiators or tuned feeders. In this case, the swr indicator belongs between the transmitter and the antenna-tuning unit so it can be adjusted to present the proper load to the transmitter. This can bring about a potentially dangerous condition. If the tuning is done for minimum swr, then the tuning unit may be mistuned so far that little power is transferred to the antenna. The same possibility holds when adjusting for maximum forward power. Only frequent checks of both will insure proper adjustment. Reflectometers with two separate meters are very helpful in this regard, as are instruments which can indicate a combination of forward and reflected powers.

I hope that this article will dispel some of the apparent mysteries surrounding directional instruments and will lead to their better utilization and understanding.

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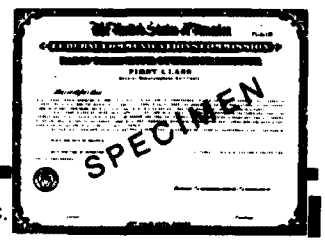
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meet the neat mnemonic

Don't let the amateur exams get you down – try these memory joggers for the basics

Alfred Wilson, W6NIF, 3928 Alameda Drive, San Diego, California 92103

There are a few individuals among us who are blessed with near-total recall. No event, number, face, name or whatever, seems to elude them. If you happen to be one of these fortunate people, this article isn't for you. But if you're like I am, then you need some kind of crutch to help you remember abstruse information which is absolutely essential in accomplishing a specific objective. The objective in this case is passing the amateur written examinations.

An interesting thing about memory is its specialization aspect. You might remember certain kinds of information extremely well and other kinds poorly or not at all. The trick to increasing your memory efficiency is to devise some method of coding the information you have trouble recalling in terms of information you can remember well. Such a device is called a mnemonic (pronounced with silent "m"). It is used by logicians, mathematicians, psychologists—even by digital computers.

There's a subtle trap lurking in mnemonic systems, though. When used in limited fields of high specialization they are quite useful. However, if applied to generalized situations, mnemonic coding tends to fall apart because of its unavoidable complexity and ambiguities. As a matter of fact, it's possible to devise mnemonics so far fetched as to defeat their very purpose: you can get to a point where you can't even remember the mnemonic for the mnemonic you're trying to remember (a bad joke, but my own). Used with care and a little thought, however, the system does work for special cases.

Consider now the newcomer to amateur radio. How can he possibly remember the deluge of equations, rules, concepts, and relationships required to pass the amateur written test? "Study," you say, and right you are! But how about helping him over the rough spots—at least until he gets to the point where, finally, the light begins to dawn?

I recall when I was first exposed to the mysteries of radio theory. (We didn't call it electronics in those days.) I took one look at the equation for resonance in a tuned circuit and thought, how am I ever going to remember this stuff? Then I flipped over a few more pages in the book and saw the equation for a damped wave (that's what comes out of a spark transmitter, youngster). I shuddered, buried the book under a stack of Captain Billy's Whiz Bangs and old Liberty magazines, then took off to see my old pal Ed, W6NHZ.

Ed was really a very astute fellow. After I explained my problem, he came up with some real help—in the form of mnemonics; only he didn't call it that. I doubt if the word was even in existence then. Anyway Ed said, "For openers let's look at Ohm's law for dc. The law was invented by George Ohm sometime in the early 1800's. If you can

remember this basic equation you can solve almost any problem involving dc circuits. But don't ask me why old George assigned the symbol 'I' instead of 'C' to represent current. Possibly he did it to avoid confusion with capacitance, whose symbol is C."

I said, "Well, the symbols for voltage and resistance I can understand, because E represents electromotive force, and R is obviously for resistance. But why 'I' for current?"

Then Ed said, "I really have no idea. Maybe George had trouble remembering the equation himself. For one thing, you'll notice that E, I and R are in alphabetical order. Perhaps he chose this sequence of symbols in an attempt at helping to remember their relationship."

I said, "Okay. All this is fine, but meanwhile I gotta memorize all this junk so I can learn why. Incidentally, why **does** E = IR anyway?"

Ed said, "Because it's the law, boy. Don't you obey the law?"

I said, "Sure, but how am I going to obey it if I can't remember whether E = IR, or I = RE, or R= EI?"

And now we come to the crux of the whole story.

"Okay," Ed said, "If you don't like E = IR, how about temporarily changing I to A (for ampere)? If you can't remember E = AR then you may as well give up. E =AR . . . get it? EAR? Them big flappy milk-wagon doors you hear with, old buddy."

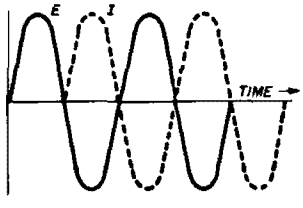
Then I said, "Yeah, yeah—I get it! In order to hear it with my ear, it's got to be changed to dc, right?"

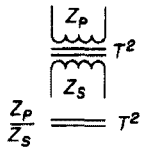
I won't repeat here what Ed's reply was, but he got his point across, and for that I'll be eternally grateful.

So I went home and stared long and hard at all the equations, symbols, and other oddments that make up the basics of our hobby. I then devised a set of coded symbols to help me remember the necessities.

Some of the more basic mnemonics are listed below with their symbolic derivations. The list is by no means complete because of space limitations. However, enough material is included to get you started; then you can invent your own. I've also added some material on transistors to keep up with the changing world.

My mnemonic symbols suit me just fine, because they're personalized. Perhaps you have better ideas to adapt the mnemonics to your particular requirements. In any event, if I've provided no more than the initial push to help a new amateur get started, then this piece will have served its purpose.

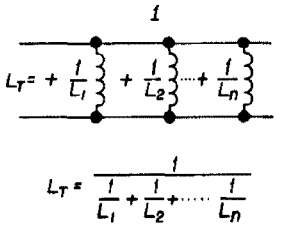
	problem	mnemonic
	**1.** Phase relationship between E and I	E is the reference. Across a pure inductance E leads I; across a pure capacitance E lags I. For X_L and X_C think: "$+X_{Leads}$ and $-X_{Comes}$ after."
$Z = \sqrt{R^2 + (X_L - X_C)^2}$	**2.** General equation for impedance	It's assumed you know Z contains R, X_L, and X_C. Impedance equals the square root of R squared plus the net reactance squared. Think: "Square **R**oot of **R** squared plus (reactance) squared." From problem 1, $+X_L$ leads; X_C comes after. This takes care of the second term under radical.



3. Impedance ratio of transformer windings

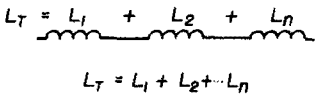
The transformer **primary** is the reference. The symbol resembles the equation. Let the equal sign represent the transformer core.

parallel and series elements



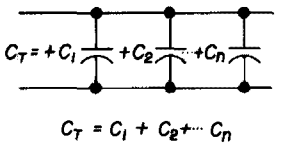
1. Combination of inductances in parallel.

Inductance and the numeral are the reference for parallel elements. The symbol resembles the equation. Think: "One over one over one Lth, plus one over one Lth, plus . . ."



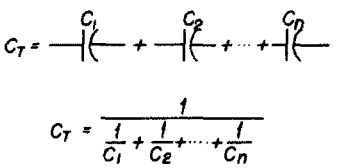
2. Combination of inductances in series

The symbol resembles the equation. Think: "String of coils, L_1 plus L_2 plus . . ."



3. Capacitors in parallel

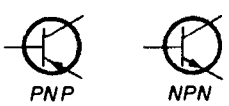
This is the exact opposite of the symbol/equation relationship of the reference mnemonic (inductance).



4. Capacitors in series

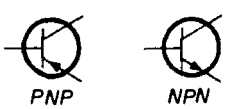
Same rationale as in problem 3

transistors



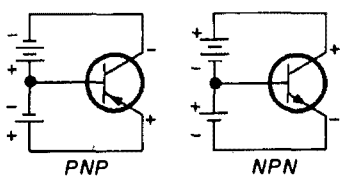
1. Direction of emitter arrow

The transistor **base** is the reference. For PNP, think: "Arrow **P**oints toward base:" **P** NP. For NPN, think: "Arrow does **N**ot **P**oint toward base:" **N** PN.



2. Element names

The emitter is the reference. Think: "Arrow emits" current from, or toward, the base. The other elements are then obvious by deduction.



3. Emitter and collector polarities with respect to the base

The polarities **must** be opposite. The **emitter** is the reference. For **P**NP, think: "Emitter **P**ositive." For **N**PN, think: "Emitter **N**egative." By deduction, the other polarities are obvious.

ham radio

photographic illustrations for amateur radio magazines

Try these suggestions when preparing photos to go along with your manuscript

William G. Blankenship, Jr. WA4GNW/W1RDR, 6709 Green Meadow Court, Louisville, Kentucky 40207

The mass circulation, high quality magazines published in this country use photographic illustrations in great quantity. These photographs, especially product illustrations, are made by skilled professional photographers who work from layouts supplied by advertising agencies, the magazines, or the client advertisers. It is customary to ship the products to be photographed, at considerable expense, to the photographic studio chosen for this work. The photographer's fee is usually quite high. The manufacturer of the product pays for magazine space, agency fees, photographs, etc.

The amateur radio magazine editor's situation is different. He depends, in most cases, on the authors of technical articles for photographs to illustrate the subjects. Most of us have noticed otherwise excellent articles in some amateur radio magazines accompanied by mediocre photographs. The editor often has no choice but to accept the photos furnished by the author. He has deadlines to meet, space and layout problems, and must also be very cost conscious. The purpose of this article is to aid the amateur photographer in making photographs of acceptable quality for publication, using his own photographic equipment, and without the use of elaborate lighting setups.

It is of course impractical, due to space limitations, to delve deeply into the technical aspects of photography in an article such as this. Many different types of cameras are owned by amateurs. It is quite obvious that a view camera is ideal for illustrative photography. This type of camera makes possible such things as perspective correction, image size control, maximum control of depth of field, etc. It is also true, however, that few hams own this type of equipment. If you do, by all means use it. Regardless of equipment

or technique, photographs for publication must meet certain minimum standards:

1. They must be **sharp.** A blurred or out-of-focus photograph cannot be improved in the making of the halftone negative for the printing plate.
2. They should be on glossy paper, with a uniform gloss over the entire print. Matte-finish prints or color prints should not be submitted.
3. They should have a normal range of contrast—the lightest and darkest areas of the print should show good detail. If this guideline is followed, a slighlly "flat" print, or one with excessive contrast, can be corrected in the plate-making process.
4. The print should be somewhat larger than the illustration will appear in the magazine. About 5x7 inches is a good size, but smaller sizes can sometimes be used.
5. Underchassis or other views showing component placement and wiring should be lighted so that confusing shadows are eliminated.
6. Perspective distortion should be minimized. For example, a high-angle shot of a unit's front panel, made with a standard camera, will cause the bottom of the panel to appear narrower than the top. This effect is due to a diminishing perspective of the vertical lines, sometimes called "keystoning."

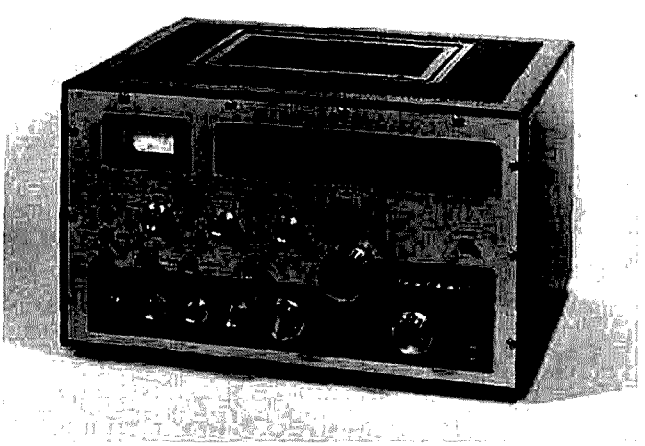
We will cover briefly some of the more important aspects of photo illustration including background, lighting, equipment and film.

background

The background against which a unit is photographed is of prime importance. The background most favored by professionals is white, seamless, heavy paper, available at photo supply or display supply stores. I normally use rolls of paper 9 feet wide and



fig. 1. In photo A, below, the vertical part of the background is reflected on the top of the transmitter cabinet; the dark cloth in photo B, to the right, eliminates the reflection.



10 yards long. However, some display stores stock 28x44-inch sheets of the same material which is more practical for the occasional user. For photographing a light colored chassis or other object, a somewhat darker background should be used. For some subjects, a

The author has been a professional photographer for many years. To lend authenticity to the methods detailed in the article, all photographs were made under the same conditions that might be found in the homes of amateurs who have an interest in photography. Commercial studio facilities were not used, and all film developing and printing was done by a custom photofinisher, as suggested in the article. **editor**

black background is ideal. The background should be large enough to extend well beyond the subject on all sides so that the editor can crop the picture to fit his format. When positioning a unit on a white background, first wipe the rubber feet with a damp cloth, followed by a dry one, to prevent "footprints" on the background.

It will be found that the vertical portion of a white background behind an equipment cabinet will sometimes cause an undesirable reflection in the cabinet top. This is illustrated in **fig. 1A.** A dark cloth, hung on the background so that its bottom edge is just outside the picture area, as shown in **fig. 1B,** will eliminate this reflection. This is a purposely exaggerated example but serves to illustrate the point. The cloth must be cropped out in the published picture, so make sure it doesn't take up too much of the background.

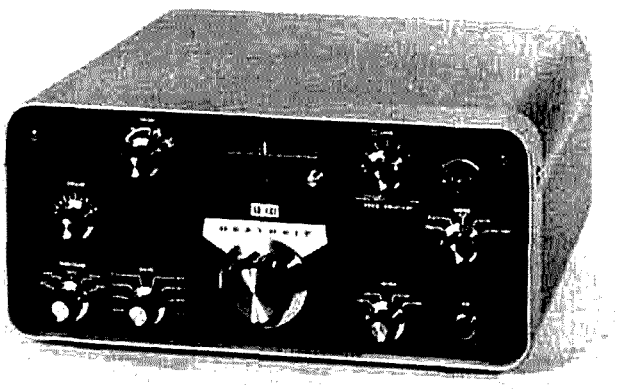
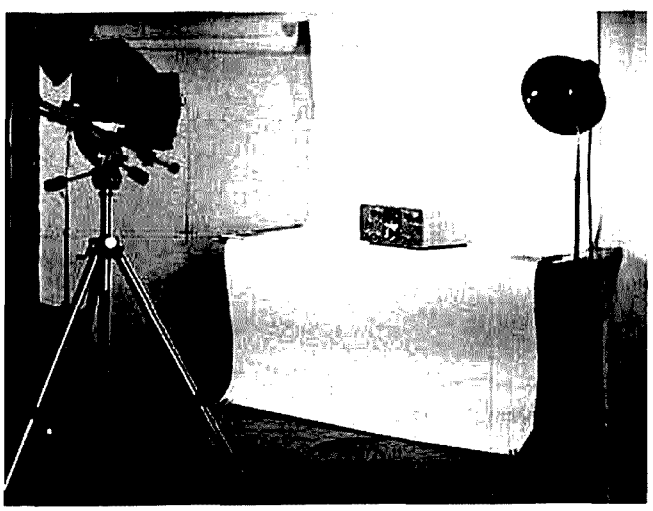


fig. 2. In this view-camera shot (right), all three planes of the unit are shown in correct perspective. The set-up is shown fixed photofloods provide lighting.



lighting

Photoflood lamps in reflectors are the best all-around light source. Reflector-floods (lamps with built-in reflectors) are satisfactory. Floodlamps make it possible to study the lighting before exposures are made, thereby avoiding specular highlights (hot spots) on reflective surfaces. Ideally, a high wattage spotlight would be used as a key light and floods would then be used as fill-lights. A study of advertising photos will reveal this widely used technique. However, this equipment is not available to the average ham so a good rule to follow when using floods is to use the minimum number necessary to do the job. This will minimize multiple shadows from knobs and other protruding parts of the unit being photographed. Refer to **fig. 2** for a photo made with a 4x5 view camera. The simple lighting set-up is also illustrated in **fig. 2.** Camera "swings" and "tilts" have been used to eliminate perspective distortion.

If photofloods are not available, a single low-wattage lamp can be used to make overall views, using the light-painting method discussed in "underchassis lighting."

underchassis lighting

A good technique for lighting cluttered areas to eliminate confusing shadows is to "paint" with light. With the camera on a tripod, focused, and the chassis properly positioned, the lens diaphragm is stopped down to the smallest opening, usually f.16, f.22, or on some cameras, as small as f.45. Incidentally, when focusing on a unit which

has considerable depth, the correct point on which to focus is about one-third of the distance from the front to the back of the unit. This is because the depth of field of a lens (the nearest and farthest areas which will be

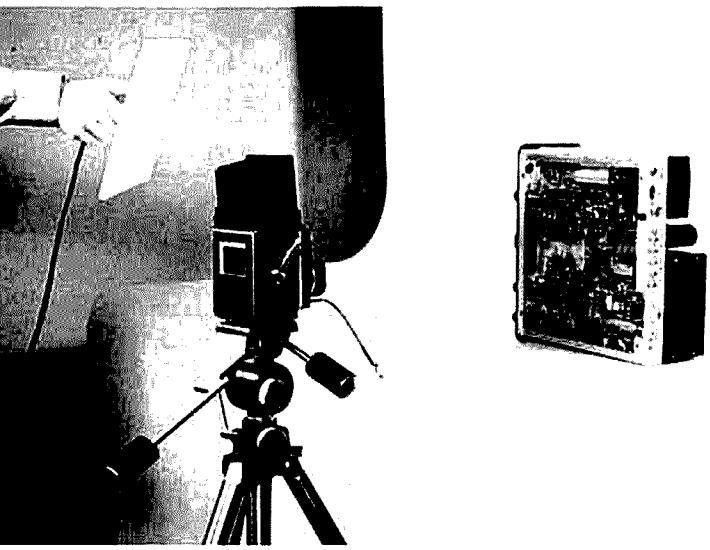


fig. 3. Simplest method for "painting with light."

in sharp focus) increases as the lens is stopped down more rapidly on parts of the subject which are farthest away and less on areas which are closest to the camera.

Set the shutter to **time** or **bulb.** In the latter case, a locking type cable release will be required. A low wattage (about 60 to 100 watt) lamp in a reflector at the end of an extension cord is used. If a reflector is not available, an 8-inch square of ¼-inch styrofoam makes a good reflector and eye shield, or a piece of white cardboard may be used. See **fig. 3** for a typical set-up.

Before opening the shutter, move the light to various positions to determine whether any areas of the subject have undesirable reflections. Avoid these positions of the light when making the exposure. Now, open the shutter and paint the light on the subject, using side-to-side, up and down and circular motions, keeping the light constantly in motion. Do not allow light to strike the camera lens; a lens hood is recommended. Room lights should be turned off.

The exposure time will depend upon lamp wattage, film speed and other factors. Typically, using Plus-X film and a 60-watt lamp, an exposure of about 3 to 4 minutes at f.22 should be about right. Subject brightness, lamp-to-subject distance and over-all technique will introduce some variables. A little experimenting will quickly give you the feel for this type of lighting. The basic reason for painting with light is, of course, to keep the shadows of wires and components mov-

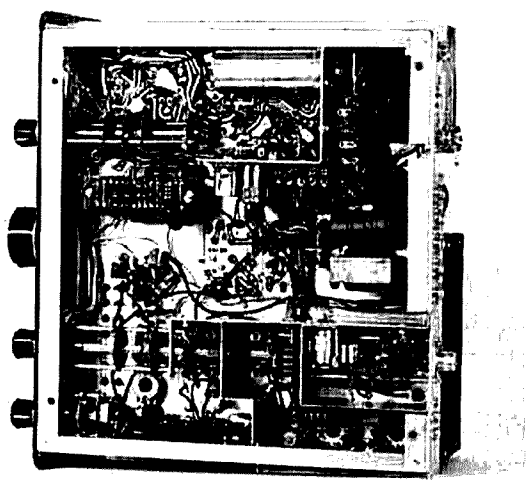


fig. 4A. Underchassis view made with fixed lights. Not bad, but "painted light" in fig. 4B is less confusing.

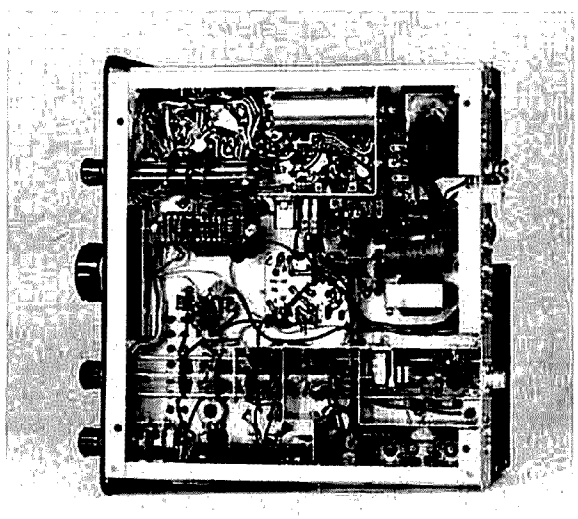


fig. 4B. Painted with light using a low-wattage bulb as discussed in the text.

ing so that no shadow will stay in the same place long enough to register sharply on the film—thus shadowless lighting. A fixed photoflood-lighted shot is shown in **fig. 4A** and for comparison, a painted-light photo made with a 60 watt lamp, in **fig. 4B.**

lighting by bounce flash

I have "painted" hundreds of commercial photographs where cluttered areas dictated such an approach, and this method is highly satisfactory. Another technique which gives good results on some subjects is to bounce the light from a flashbulb. For example, a chassis standing on edge or placed flat on the floor with the camera above it can be well lighted by this method. A large white card or movie screen positioned so that the light from a flashbulb can be bounced onto the subject will diffuse the light and produce a near-shadowless shot of the subject. Open-flash technique (open shutter, fire bulb, close shutter) is recommended. Camera diaphragm setting should be about two stops larger than if direct flash were being used. Direct flash on this type of photo is, of course, contraindicated. Moving the screen to two or more positions and bouncing the light from a flashbulb in each position will result in more even lighting. A bounce-light arrangement is shown in **fig. 5B.** The unit shown in **fig. 5A** was photographed by bouncing the light from three flashbulbs, each with the screen in a different position.

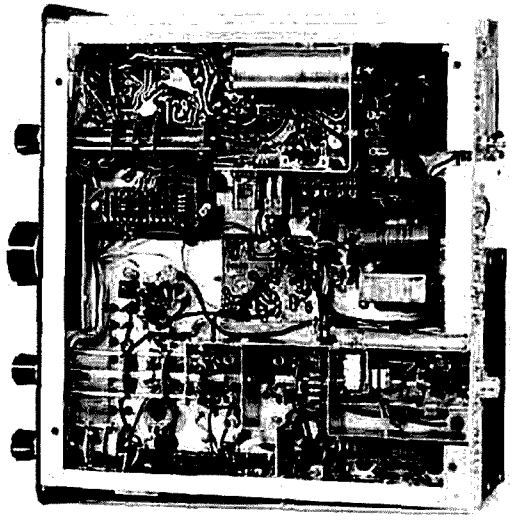


fig. 5A. Undorchassis shot made with bounce flash.

perspective correction

The transmitter in **fig. 2** was photographed with a 4x5 view camera. The front panel, right side and top are shown in correct perspective. A lens of 10-inch focal length was used. When using fixed lights, an exposure meter should be used. A $2^1/_4$x$2^1/_4$ twin-lens-reflex camera with a normal (80 mm) lens, and without the swings and tilts of a view camera, was used for **fig. 6A.** Perspective distortion is quite evident. This set-up is shown in **fig. 6B.** The improved perspective achieved, using the same camera, by shooting from a lower angle (keeping the film plane of the camera parallel to the transmitter's front panel) and from a greater distance, is shown in **fig. 7A; fig. 7B** shows the camera position. Thus, it is better to back up far enough to minimize perspective distortion and then make an enlargement to the desired size.

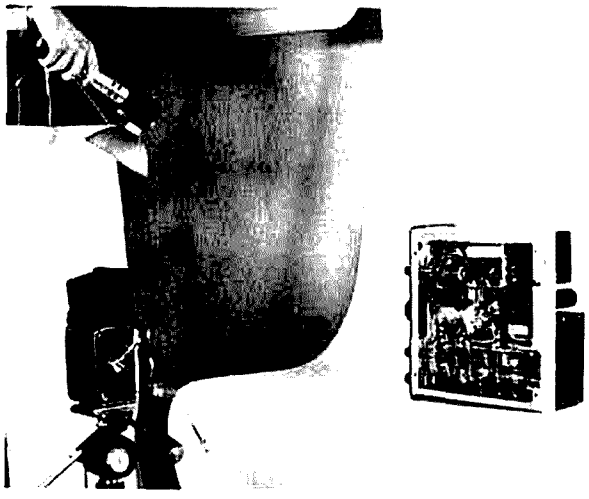


fig. 5B. Bounce-flash set-up using a movie screen.

foreshortening

In addition to perspective distortion, another undesirable effect that results from

too-close use of cameras with standard focal length non-interchangeable lenses is foreshortening. This simply means that the part of the subject nearest the lens will be disproportionately large when compared with areas of the same subject which are farther from the lens. Example: portraits made too close to the subject will result in noses too large and ears apparently set too far back. Foreshortening and front-to-back keystoning are evident in **fig. 8.** This shot was made with a 35 mm camera using a 50 mm lens. The same set-up was used for the perspective view in **fig. 9A.** Distortion is very noticeable. Camera position is shown in **fig. 9B.** Changing to a 105 mm telephoto lens, the camera was moved back to more than twice the distance (see **fig. 10B**) with the greatly improved results seen in **fig. 10A.** The image size on the film was as large as that obtained with the 50 mm lens close up in a close-up position.

some notes on equipment

Many amateurs own 35 mm cameras. Good quality photographs from the small negatives produced by these cameras are more difficult to make than those from larger sizes, such as $2^1/_4$x$2^1/_4$, 4x5, etc. However, by using fine-grain film, properly fine-grain developed and printed, high quality photographs can be made. It is interesting to note that a good 35 mm camera usually has a sharper lens than the larger cameras. The resolution of lenses used on larger cameras

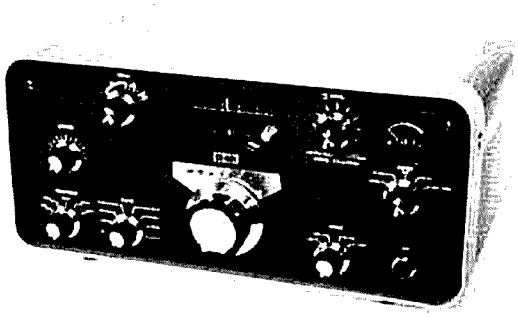


fig. 6A. Distorted view—standard camera positioned too close and too high.

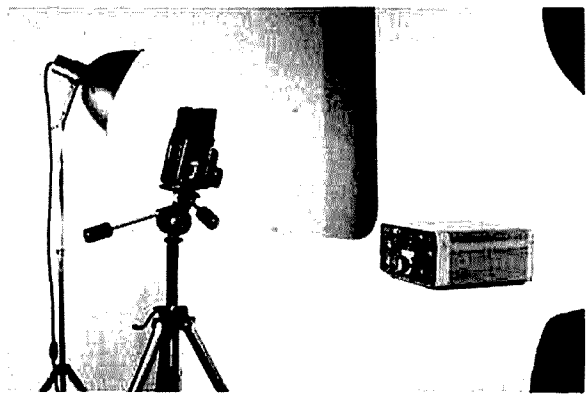


fig. 6B. Camera position used in fig. 6A; note downward tilt.



fig. 7A. Made with the same camera as fig. 6A. Improved perspective. Camera lower and camera-to-subject distance increased.

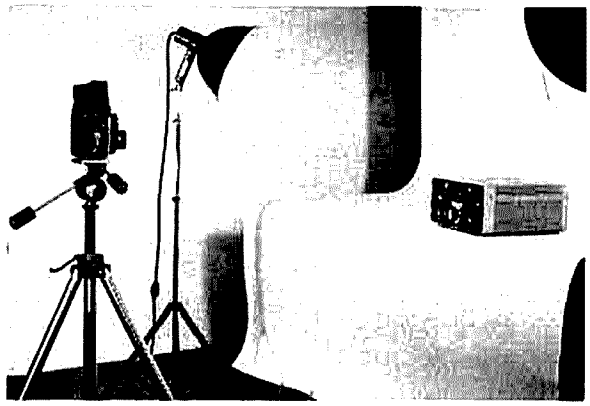


fig. 7B. Correct camera position; note that film plane is parallel with front panel of equipment.

need not be as great, in general, as those used on 35 mm. A 4x5 negative need be blown up only two diameters for an 8x10 print, whereas the same size print from a 35 mm negative requires about seven diameters enlargement. If the 35 mm camera has interchangeable lenses, a medium telephoto

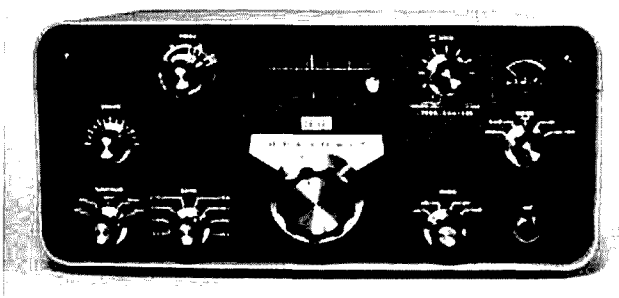


fig. 8. Foreshortening caused by camera being too close to subject; knobs appear canted outward and *downward*.

(100 mm to 135 mm) should be used when photographing equipment. I have read articles which stated, "the longer the lens the better". This is not true. A long telephoto lens invariably introduces an undesirable "telephoto effect" or flattening of the subject. Also, the depth of field of a medium telephoto will be greater at any f stop than that of the long telephoto. With any interchangeable-lens camera a lens of approximately twice the focal length of the camera's normal lens will yield pleasing results when making close-up shots of equipment.

A comment on print quality: the best 8x10 prints are made by **contact** printing 8x10 negatives. So it follows that one should, with any camera normally used by amateurs, get the largest negative image size possible consistent with good perspective as outlined in the foregoing text. This is especially true when using 35 mm.

Another camera in widespread use is the Polaroid. These cameras have the advantage of letting you see a finished print in seconds. However, most models have some inherent disadvantages so that the use of standard Polaroid cameras to photograph electronic gear is, in general, not recommended One possible exception is the use of a Polaroid back on a view camera.

parallax

Some cameras have viewfinders which see a different field of view than that seen by the lens. The difference in these fields of view is called parallax. On distant subjects, this is unimportant. In close-up photography, compensation must be made. You have probably seen photos of people with tops of their heads cut off. Someone didn't take parallax into account! In close-up work, center the unit to be photographed in the viewfinder. Then, before shooting, move the camera up and possibly sideways, depending upon the location of the viewfinder, until the lens occupies the same position.

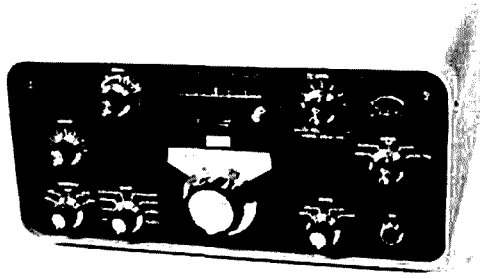


fig. 9A. Made with a 35 mm camera, 50 mm "normal" camera.

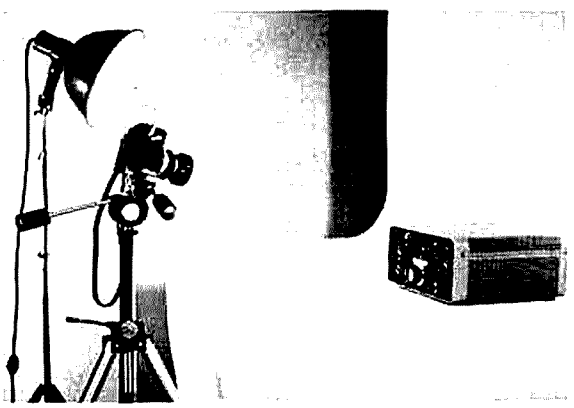


fig. 9B. Camera position used for taking fig. 9A. Good image size but the camera is much too close.

film and developing

Most camera users probably have a favorite film. For what it is worth, I have found Super Panchro Press Type B to be an excellent all-around sheet film for commercial subjects. For 35 mm cameras, Plus X is recommended. Panatomic X is somewhat finer-grain, but Plus X, properly fine-grain developed, yields a somewhat more brilliant print with no objectionable graininess. For cameras using 120 film, Professional Plus X is a good choice.

If you don't do your own processing, try to avoid using "drugstore" photofinishing. This grade of work is generally not suitable for good reproduction. Most cities have custom photofinishers who do high quality work for professionals. Use their services, if possible; the additional cost is justified.



fig. 10A. Made with 35 mm camera, 105 mm lens. Longer lens provides same image size as fig. 9A, but perspective is improved.

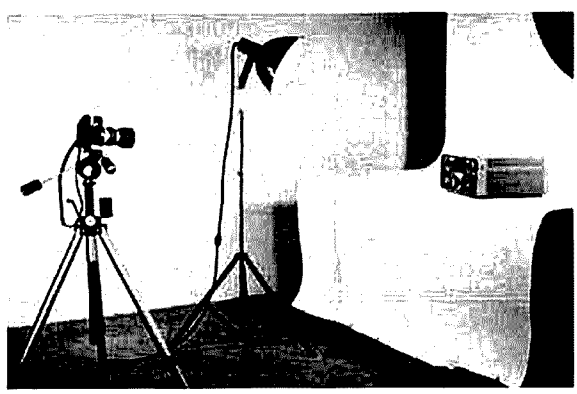


fig. 10B. Camera position used for taking fig. 10A. Camera is slightly lower and film plane is parallel with the front panel.

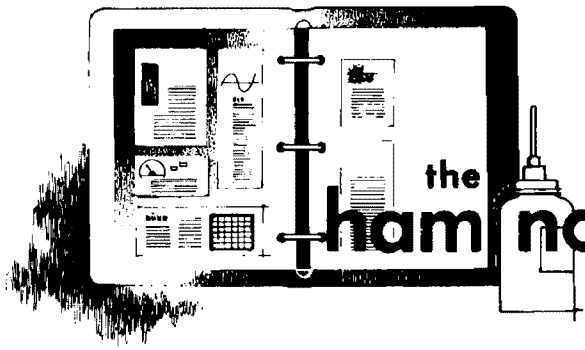
summary

To summarize, if you don't have a view camera, shoot as nearly level with the unit as possible (film plane parallel with front panel) to avoid perspective distortion. Incidentally, some degree of correction of converging lines can be made by tilting the easel when enlarging. Dimensions of the unit photographed can be indicated in the article. A rule may be included in the picture if desired, but place it so it can be cropped out if the editor doesn't want to use it. Place the light or lights so that pleasing illumination is achieved and hot spots are minimized on reflective surfaces. Paint or bounce the light into cluttered areas for greater clarity. Get far enough from the subject to minimize foreshortening. It is much better to have a smaller, distortion-free image, which can be blown up, than it is to have a larger one that overemphasizes areas that are closest to the lens. Finally, make sure that the pictures you submit are **sharp.**

If you write descriptive matter for your photos to point out specific areas, do not in any case write on the photos. Instead, you should: a) write on a separate piece of paper and attach it flush with the bottom edge of the photo, using cellophane tape as a hinge; b) attach a tissue overlay to the print and write **lightly** with a soft pencil to prevent scoring the print, or c) send a duplicate print on which you have indicated the areas to be emphasized.

Here's a final item often overlooked by amateurs who submit photos to editors. Your high quality pictures are useless if they arrive at the editorial offices bent and cracked. Make a sandwich for your photos by cutting two pieces, large enough to extend at least a half-inch beyond the edges of your prints, from a corrugated carton. Tape the edges of the corrugated board so the photos can't shift in handling. Most photo supply houses also carry photo mailers which are ideal for the purpose. Happy shooting!

ham radio



the ham notebook

32S-1 cw modification

The 32S-1 as originally designed uses a 2.1-kHz-bandpass mechanical filter centered on 455 kHz and two bfo crystals, one above 455 by 1350 Hz and one 1350 Hz below, operating on 456.350 and 453.650 kHz respectively.

In cw operation, a tone oscillator at 1350 Hz is fed through the mike amplifier, balanced modulator and mechanical filter to create a cw carrier. The carrier is at 455 kHz in the low i-f by virtue of the relationship of the bfo crystal frequency (456.350 kHz) and the modulating frequency of the tone oscillator (1350 Hz) so that 456.350 − 1.350 = 455 kHz.

The quality and stability of the diodes available at the time the 32S-1 was designed and manufactured was far below those available today. It is difficult to maintain carrier suppression better than 50 dB over a long length of time with the original components and circuitry. However, a very easy alternate method of vastly increasing carrier suppression is possible.

A 457.55 kHz crystal* is stocked for those amateurs who operate RTTY using a 2125/2975 tone keyer; this places the audio response between 1500 and 3600 Hz. Of specific importance is the fact that the crystal is removed from the skirt of the mechanical-filter bandpass by 1500 Hz rather than the normal 300 Hz, so an additional 40 to 50 dB carrier suppression is obtained. The 1350 Hz tone oscillator no longer puts the "carrier" at 455 kHz, and its second harmonic could pass through the mechanical filter with little attenuation. Therefore the tone frequency must be raised to, ideally, 2550 Hz (457.55–2.550 = 455 kHz).

Refer to the 32S-1 schematic diagram. Locate switch S8F, just below V2B, the 6U8A bfo tube, and to the right of the tone oscillator. The upper sideband crystal, Y15, is shown wired to the most clockwise position (cw); a jumper goes to the adjacent contact, which is the usb wiper, and to the **tune** and **lock-key** wipers. The jumper should be removed so that Y15 is connected to the usb wiper only. The jumper from the **tune** and **lock-key** wipers should be connected to the cw wiper. Do not solder yet, as the new 457.55 kHz crystal should be connected between the cw wiper and pin 3 of V2B (the pin where the other crystals are connected).

To raise the frequency of the tone oscillator, C111 should be changed from its present value of 430 pF to 220 pF, and C110 should be reduced from 390 pF to 180 pF. The exact frequency of 2550 Hz might not be realized but the resultant frequency is sufficiently close to meet requirements.

Those amateurs who have operated modified 32S-1's report that they don't object to the higher pitched sidetone, although many use the sidetone from their electronic keyer so the frequency difference is not a consideration. After this modification, carrier suppression should be on the order of 80 dB. This is sufficiently below the white noise output to be undetectable. On-the-air reports have been excellent.

* The 457.55 kHz crystal is available from the Collins Radio Company, Cedar Rapids, Iowa 52406. Order part number 290-8708-000; $5.50 plus sales tax, postage and insurance. The price is subject to change at any time without prior notice.

painless vox keying

The convenience and speed of vox keying is well known by those who have this feature in their transmitter or transceiver. With this method, changing from receive to transmit mode occurs at the first touch of the key, which activates the vox circuit normally used on phone. A recently purchased transceiver, while excellent in many other respects, didn't have this feature. So I incorporated the keying method shown in **fig. 1** to provide vox keying. It also has other possible uses.

Before using this method with your equipment, you should determine what additional modifications are necessary. The following conditions must be met.

1. The speech amplifier must be disconnected from the vox input or its power supply when the transmitter is placed in the cw position.

2. The vox circuit must remain in operation with the transmitter in the cw position.

3. The vox relay must control the same circuits as the cw operate switch.

If all these conditions are met, all that's necessary is to connect the keyed audio to the microphone input, place the operate switch in vox position, and go ahead.

Grid block keying was used in the original circuit. A −180 V line was used to cut off the driver stage during standby. The −180 V line was shorted directly to ground by the key. This made it difficult to use the same contacts to operate other circuits. In **fig. 1,** a silicon controlled rectifier replaces the key contacts, and a positive-gate turn-on voltage is keyed. This removes the shock hazard and provides a keyed source of low voltage to turn on other circuits.

The audio oscillator module is a small code-practice type available at most amateur equipment dealers. The speaker is a tiny 8-ohm transistor radio replacement. Your junk box might provide a phase-shift or unijunction oscillator.* However, at the price of the code-practice module, constructing one would hardly be worth the effort.

R. J. Beck, VE7IG

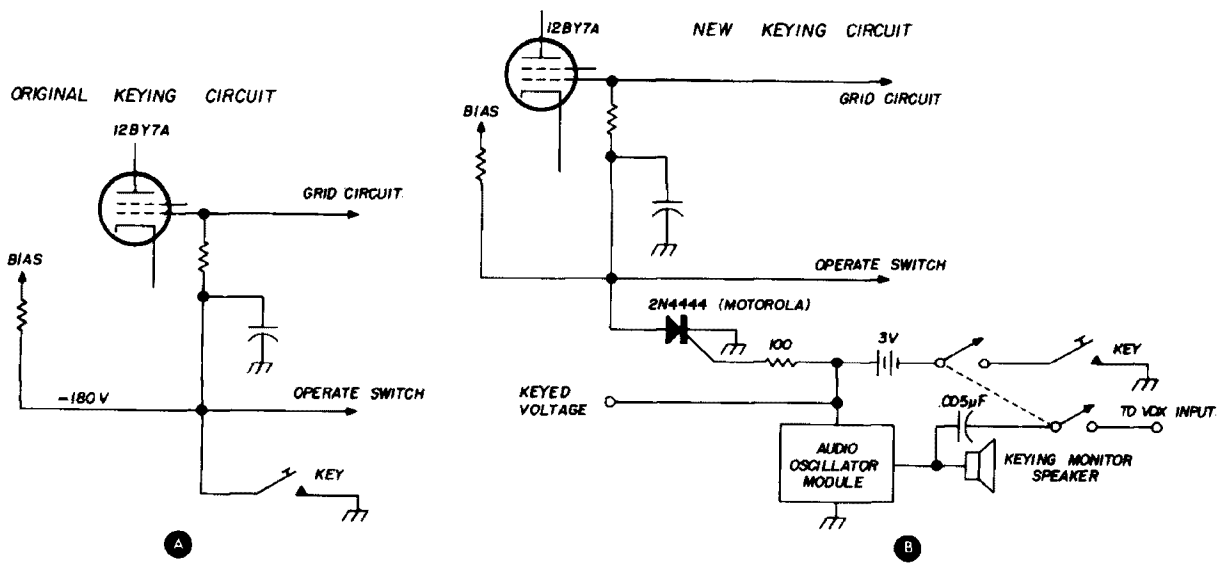
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Don Farrell, W2GA

* E. R. Davisson, K9VXL, "Replace One Unijunction Transistor with Two Transistors," **ham radio,** April 1968, pp. 58-59.

fig. 1. Original vox keying circuit, A, and new circuit, B.



update your Swan 250

The new Swan 250C has quite a few improvements over the older model 250. These include higher gain and better selectivity. Strong signals are handled much better by the new model. You can improve these characteristics in your 250 Model with a few simple circuit changes. You can also easily install the Swan noise blanker.

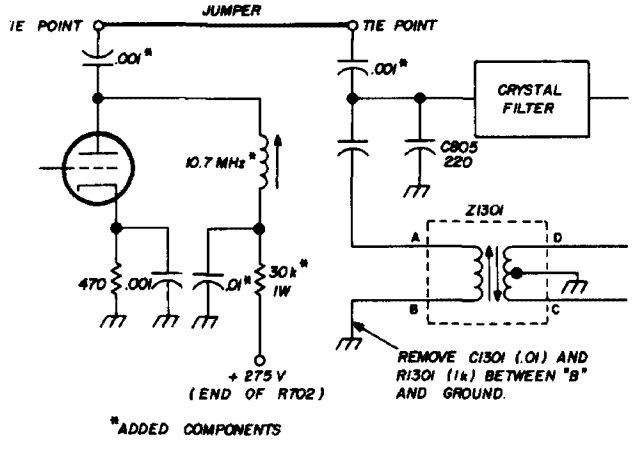


fig. 2. Circuit changes for higher gain and improved selectivity in the Swan 250.

Fig. 2 shows the circuit modifications. Remove the top and bottom covers, and remove the 2-inch wire from pin 5 of V7 (6HA5). Remove resistor R1301 (1 kilohm) and capacitor C1301 (.01 μF) from terminal "B" of Z1301 balanced modulator transformer. Ground terminal "B". Add a single tie point near pin 5 of V7 and one near terminal "A" of Z1301. Connect a 0.001-μF capacitor from one tie point to pin 5 of V7; install another 0.001-μF capacitor from the other tie point to the input of the crystal filter. Install the slug-tuned coil, the 0.01-μF bypass, and 30 kilohm resistor in the plate lead of the 6HA5 as shown in the schematic.

Add a short jumper wire between the two new tie points you installed next to V7. If you wish to install the Swan noise blanker later, you can remove the jumper and put the blanker between these two points. When all the new parts are installed, align the front end and i-f as outlined by the manuals. Be sure to carefully peak up the new 10.7-MHz coil and the slugs in Z1301.

Harold D. Mohr, K8ZHZ

miniature sockets

Many MARS members have been receiving miniature vacuum-tube sockets. These sockets have 8 holes with one empty space so it's quite evident which contact is number 1. These sockets are ideal for 6- or 8-pin integrated circuits or 6-pin transistors.

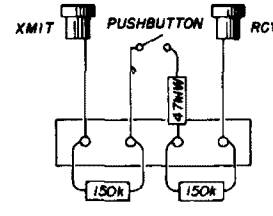
James T. Lawyer

spot switch for the heathkit sixer or twoer

A momentary pushbutton switch and a one-watt, 47k ohm resistor are all that are needed to put a spot switch on the sixer or twoer.

The resistor is in series with the switch, and they are mounted across the two middle terminals of the terminal strip behind the neon lights. **Fig. 3** shows the connections.

fig. 3. Spot switch for Sixer and Twoer. Pushbutton switch and 47k resistor are mounted across two middle terminals of strip behind neon lights.



This puts about 50 volts on the B+ line to the driver stages of the transmitter when in the receive position. This is enough to get a carrier indication. The final doesn't draw current, because the cathode is open until the transmit switch is operated.

The resistor drops 210 volts at a measured 5 mA. The dissipation is one watt. Because it's not on for any length of time, I used a one-watt resistor without allowing for safety factor.

Edmond DeCloedt, WA6FNR

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WANTED by FBI



BENJAMIN HOSKINS PADDOCK

BENJAMIN HOSKINS PADDOCK, also known as Perry Archer, Benjamin J. Butler, Leo Genstein, Benjamin Hoskins, Jr., Ben H. Paddock, Jr., Benjamin Hodgens Paddock, Benton Hoskins Paddock, Pat Paddock, Patrick Benjamin Paddock, "Big Daddy," "Chromedome," "Old Baldy". FUGITIVE, ESCAPED FEDERAL PRISONER. IDENTIFICATION ORDER NUMBER 4261.

Benjamin Hoskins Paddock is being sought by the FBI for violation of the Escape and Rescue Statute.

On December 31, 1968, Paddock escaped from the Federal Correctional Institution at La Tuna, Texas.

A Federal warrant for his arrest was issued on February 3, 1969, at El Paso, Texas. Paddock has been arrested numerous times for bank robbery. He has been convicted of bank robbery, automobile larceny and confidence game.

Paddock's personality has been diagnosed as being psychopathic, egotistical and arrogant. He is reputed to be a very smooth talker, an avid bridge player, and a stock car and baseball enthusiast. Paddock has been employed as an automobile mechanic, electrician, promoter, salesman, service station operator, clerk and welder. He was formerly licensed as K7JIH and might be known to readers of this magazine. Paddock has utilized firearms in the past and should be considered extremely dangerous.

DESCRIPTION OF BENJAMIN HOSKINS PADDOCK

Born, November 1, 1926, Sheboygan, Wisconsin. Height, 6'4". Weight, 245 pounds. Build, large. Hair, blond, balding (head may be shaved). Eyes, gray or green. Complexion, medium. Race, white. Nationality, American. Scars and marks: scar over right eyebrow, vaccination scar upper left arm, scar over right knee, birthmark right ankle. Remarks: may wear glasses or contact lenses. Social Security Number used: 340-18-9361.

If you know anything about this man contact your local FBI office.

ham radio

index to volume II–1969

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